THE UNIVERSITY OF ALBERTA

SEMICONDUCTORIZED HIGH VOLTAGE SUPPLY

A THESIS

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by

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ABSTRACT

This research consists of the development of a completely semiconductorized, 1000 volt, well regulated, D.C. power supply.

Because transistors cannot handle voltages of the magnitude of 1000 volts, the conventional series or parallel regulator schemes could not be applied. The system developed here makes use of a generated sinusoidal signal which is amplified, then transformed to the required voltage after which it is rectified and filtered. The D.C. output voltage is compared against an accurate reference and the error signal is used to correct the amplitude of the generated signal.

The model that was built performed according to the specifications set, although the transient response was rather slow.

ACKNOWLEDGEMENTS

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The writer wishes to thank other members of the Department for their co-operation and helpful suggestions.

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CHAPTER I

INTRODUCTION

The subject of this research was the development of a 1000 volt, well regulated, power supply. The specifications assigned to this project are listed below:

1. Only semiconductors were to be used.
2. The output voltage was to be 1000 volts D.C.
3. The variation of load current was to be from 0 to 10 milliamperes.
4. The voltage regulation was to be 0.01 per cent or better.
5. The transient response time was to be short.

Since transistors cannot handle voltages much beyond 100 volts, the conventional series or parallel regulator schemes could not be used. A block diagram of the scheme proposed is shown in Figure 1.1.

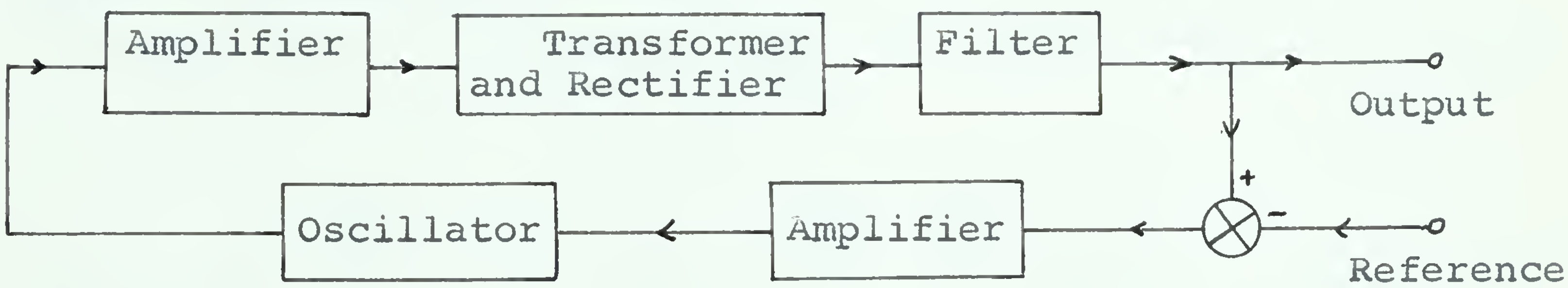


FIGURE 1.1

FUNCTIONAL DIAGRAM OF PROPOSED SYSTEM

The output signal of the oscillator was to be amplified by a power amplifier to a proper power level. To obtain the necessary voltage a power transformer was to be used to step up the voltage of the amplifier output. After the voltage was stepped up to the proper level, it was to be rectified, filtered, and used as the output of the Regulated Supply. The output was to be compared to an accurate reference, which was at a lower voltage level. A variation in the output would cause an error which would be amplified and fed back to the oscillator. This error was to be fed into the oscillator in such a fashion so that the amplitude of the oscillator output would change to oppose the error.

The frequency of the oscillator feeding the power amplifier was predetermined at 4000 cycles per second. This frequency was chosen to ease the filtering without exceeding the power transistor cutoff frequency. Because of this relatively low frequency, the response of the Regulated Supply to a step load was rather sluggish.

This was a type zero system, which means that an error must exist before the system will regulate. Therefore, the loop gain had to be made high enough so that the steady state error would be within the specified limit. The accuracy of the output was also dependent upon the accuracy of the reference, which was obtained by the use of zener diodes.

CHAPTER II

OPERATION

2.1 DETAILED DISCUSSION OF OPERATION

A full description of the system operation is given in this chapter. This includes the description and mode of operation of the components used to built up the overall system. To aid the discussion, a detailed block diagram is shown in Figure 2.1.

The output voltage of this Power Supply was compared against an accurate reference, which was obtained by the use of two zener diodes operating in series. The zener voltage of the two diodes was approximately 17 volts. Consequently, a voltage divider was connected across the system output terminals so that 17 volts, which were representative of the output, could be obtained. The values of the Voltage Divider resistors were chosen in such a way so as to draw approximately 1.1 milliamperes from the output. Included as part of the Voltage Divider was a small trimmer resistor so that the 17 volt output and the Reference Voltage could be set quite accurately to the same value when the output voltage was 1000 volts.

Having the two voltages, the next problem was to make the voltage representing the output follow the reference voltage with an error smaller than 0.01 per cent.

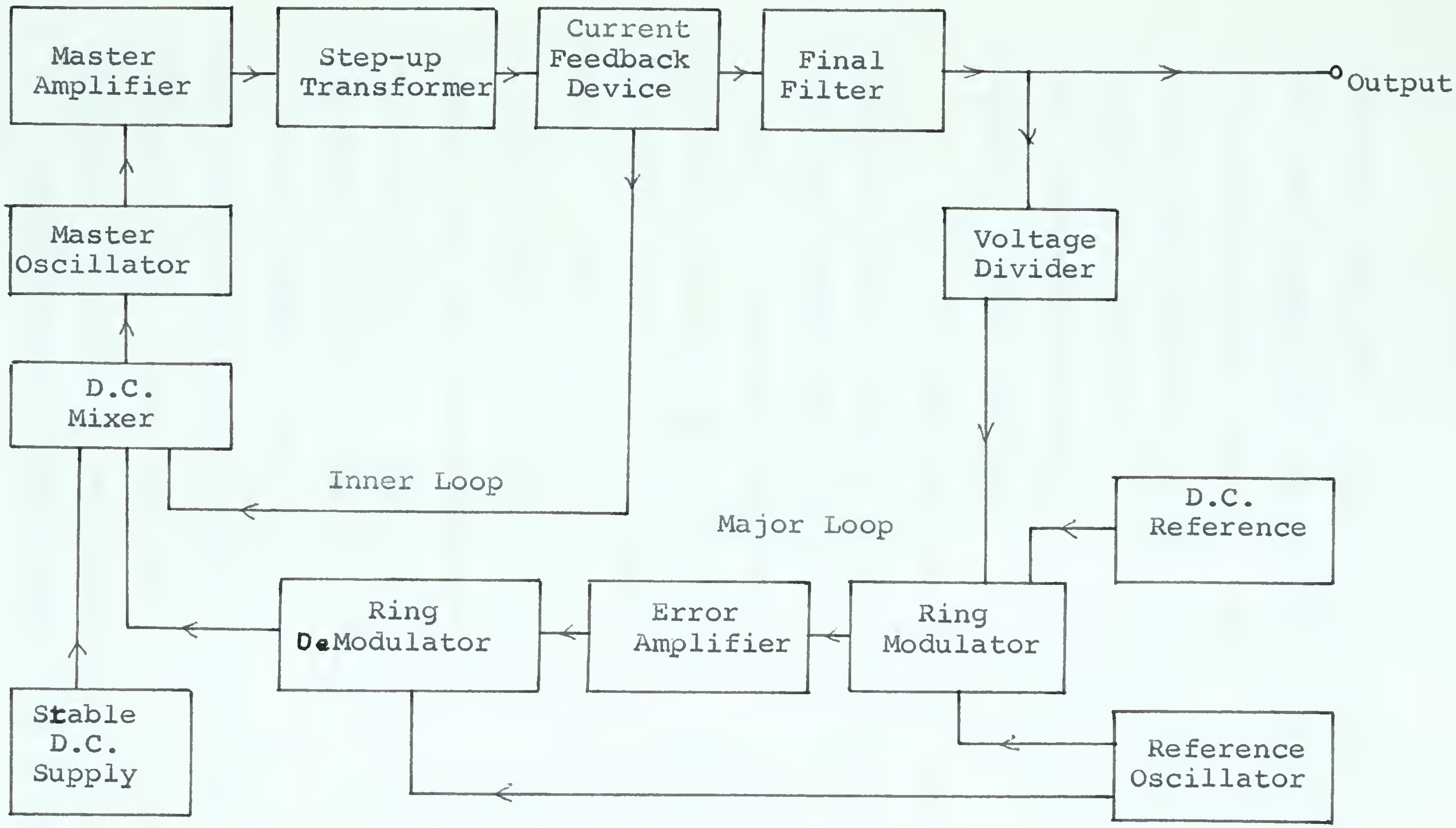


FIGURE 2.1

DETAILED BLOCK DIAGRAM OF REGULATED SUPPLY

To compare the two voltages, the negative terminals of the Reference Voltage and the voltage representative of the output were connected. The two positive terminals were connected to the input terminals of the Ring Modulator, so that if the two voltages were not exactly equal there would exist an A.C. voltage proportional to the difference across the output terminals of the Modulator.

The Ring Modulator was the most sensitive component because it had to respond to an error less than the allowable regulation, which was 0.01 per cent of 17 volts or 1.7 millivolts. Since the Ring Modulator was primarily a current operated device and the current representative of the output was approximately 1.1 milliamperes, it followed that the current to which the Modulator must respond was at the most 0.11 microamperes. It can be concluded that voltages were compared and the corresponding current, which was caused by the voltage error, was used in the correction of the output. The carrier frequency of the Ring Modulator was selected at a value of 10,000 cycles per second. This value had to be higher than the frequency of the Master Oscillator, which was 4000 cycles per second, and had to be low enough to avoid capacitive coupling in the transformers and diodes.

The Modulator output signal had an amplitude proportional to the error. This signal was amplified by the Error

Amplifier. The output of the Error Amplifier was fed into the full-wave Ring Demodulator. Since the Modulator and Demodulator were phase sensitive, a negative or positive error appeared as a negative or positive signal at the output of the Demodulator. At this point it should be mentioned that since the reference signal of the Modulator had to be set at a certain phase with respect to that of the Demodulator, they both were controlled by the same Reference Oscillator. A phase shifting device was put in front of the Demodulator so that the phase between the Modulator and Demodulator could be adjusted. The output of the Ring Demodulator appeared as a full-wave rectified signal which had an amplitude proportional to the error of the output of the Regulated Supply.

It was found that the output impedance of the Regulated Supply, with the major feedback loop open, was very high. Consequently, as the load current increased the output voltage dropped drastically. The drop in the output voltage was so large that the major loop, which was primarily built for sensitivity, did not have the range to correct the output. This phenomenon created the need for a feedback which was proportional to the load current. A convenient way of achieving this was by putting a small resistor in series with the center-tap of the Power Transformer. This was the only place where

a voltage proportional to the load current could be obtained with respect to ground potential and yet not affect the output voltage of the Regulated Supply.

A D.C. voltage proportional to the error and a D.C. voltage proportional to the load current were now available. These two signals, together with a constant D.C. voltage, were summed by connecting them in series at the input of an emitter-follower, D.C. amplifier. The polarities of these three voltages were chosen in such a way so that their sum would increase with increased load, and that it would also increase when the output of the Demodulator indicated that the Regulated Supply voltage was less than a 1000 volts. The reason that these voltages had to be added with the aid of an emitter-follower was due to the fact that a current gain was needed. Without an emitter-follower amplifier the direct current that was flowing through the Demodulator and caused by the two other sources, disturbed the operation of the Demodulator. The impedance of the emitter-follower was high enough so that the current flowing in the components at the input was small enough to avoid ill effects. Often D.C. amplifiers are undesirable because of their inherent drift. The emitter-follower was used because it is the most stable configuration and offered a small amount of drift which could be taken care of by the regulation.

The output voltage of the emitter-follower was proportional to the sum of the constant voltage, the voltage proportional to the load current, and the voltage proportional to the error of the output of the Regulated Supply. For this reason it is referred to as a D.C. Mixer. The output of this D.C. Mixer acted as a supply voltage for an oscillator which is referred to as the Master Oscillator. Since the amplitude of the output of this Oscillator varied linearly with the supply voltage, its output was proportional to the load and error of the Regulated Supply.

The output of the Master Oscillator, which was at 4000 cycles per second, was amplified by the Master Amplifier to produce a power output of about 15 watts. Since the voltage that the transistors can handle is relatively low, the amplification of the Master Amplifier was mainly current amplification. Consequently, the output voltage of the Amplifier had to be stepped up about 60 times, with the aid of the Power Transformer, in order to be able to obtain a D.C. voltage output of at least 1000 volts.

The secondary of the Power Transformer was center-tapped so that only two diodes would be needed for full-wave rectification.

As a smoothing filter, a single L-section consisting of a choke input was used. The output of the smoothing filter

appeared across the output of the Regulated Supply.

2.2 CHOICE OF COMPONENTS

For the D.C. Reference, two stable zener diodes were used. A standard cell may have been considered. However, since semiconductors were used and are essentially current controlled devices, it made the standard cell practically impossible. The reason is that during transients or when switching off, the cell may be damaged temporarily or even permanently. On the other hand, these zener diodes, which need a very stable supply to drive them in order to achieve extreme accuracy, are much more rugged.

Because a high steady state accuracy was required it was best to modulate, amplify, and demodulate the error signal to achieve D.C. amplification. The major problem of a D.C. amplifier is drift, that may be caused by power supply variation or by temperature effects on transistor characteristics and on components like bias and load resistors. Drift may cause large output signals for zero input to the amplifier so that large steady state errors may result. Also, drift may cause the amplifier to become unsymmetrical. For low gain and low stability a D.C. amplifier may be used satisfactorily with the proper stabilizing techniques, but for high gain or high stability or both it is not satisfactory.

For a modulator, the Ring Modulator was chosen mainly for its simplicity. Also, it provided zero output signal for zero input, and it had a relatively low output impedance. Its disadvantage, which was realized at high sensitivity, was that the diode pairs used across the reference and the transformer center-taps had to be exact, otherwise there was a slight output for zero input. However, this was not a major drawback. The Ring Demodulator was used for similar reasons as the Modulator.

A mechanical chopper was considered for a modulator. One disadvantage was that the frequency used was high and this made the chopper practically impossible. Also, due to the small signals to be handled this chopper would have required gold plated contacts, or the equivalent, which would have made the chopper expensive.

The Error Amplifier was a three-stage amplifier, which consisted of two emitter-follower drivers and a push-pull output. The reasons for using the emitter-follower stages were for impedance matching and for current gain. Even though the output impedance of the Modulator was relatively low, it was still high as an input to a common base or common emitter stage. The push-pull stage was used to eliminate even harmonics in the signal.

Two oscillators were built for this system: one as the Reference Oscillator for the Modulator and Demodulator, and one as the Master Oscillator which supplied the Master Amplifier. Since in both cases the outputs of the Oscillators were to be transformer coupled, it was most convenient to use transformer feedback because the transformers were hand wound.

Another major item built was the Master Amplifier which supplied the output. This consisted of three stages. Since the output impedance of the Master Oscillator was high, the first stage was an emitter-follower which was built together with the oscillator to prevent pick-up between the two components. The emitter-follower was followed by a common emitter stage which drove the push-pull output stage. The output stage was operated in class B since this provided the maximum voltage output, which was desirable. Also, the class B operation provided the optimum efficiency which is desirable when large amounts of power are handled.

The Power Transformer was built using a ferroxcube core, because this Transformer had to operate at a fairly high frequency, and ferroxcube material responds well to high frequencies. This Transformer was hand wound, therefore, it was convenient to include a center-tap so that full-wave rectification could be achieved with two diodes. Since the

diodes necessary to rectify high voltage are expensive this proved to be an economical step.

2.3 PHYSICAL CONSTRUCTION

It was convenient to build the components in separate component boxes to aid the adjustment of the separate devices. The components were terminated by the use of shielded wire and telephone jacks and plugs. Proper building techniques, which included necessary shielding, grounding, soldering of temperature sensitive components, and circuit layout, were applied. It should be pointed out that the actual building took a considerable amount of time.

CHAPTER III

DESIGN

3.1 INTRODUCTION

The designs and construction of the individual components are discussed in this chapter.

The components built are listed below:

1. Reference Supply and Voltage Divider.
2. Ring Modulator.
3. Error Amplifier.
4. Ring Demodulator
5. Reference Oscillator
6. D.C. Mixer
7. Current Feedback Device
8. Master Oscillator
9. Master Amplifier and Power Transformer.
10. Rectifier and Final Filter
11. Power Supplies.

3.2 REFERENCE SUPPLY AND VOLTAGE DIVIDER

The design of the Reference Supply and Voltage Divider was determined experimentally. The design of the raw supply which was used to drive the zener diodes is included in section 3.12. The zener diodes used for the

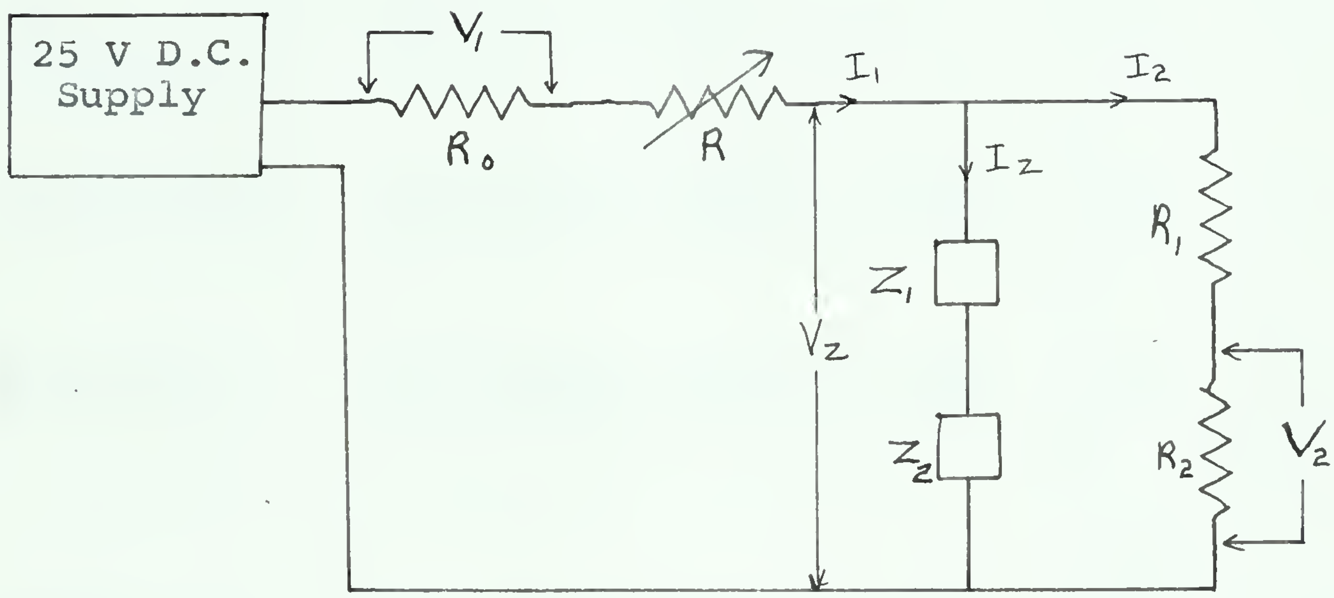
Reference Supply were the 1N430A (Hoffman). These diodes are very temperature stable. The manufacturer's specification is that the voltage stability is $\pm$0.1 per cent or less over the temperature range -55°C to +100°C. In order to achieve this result consistently, the current through the 1N430A diodes should be held to milliamperes $\pm$1.2 per cent.

A model of a regulated power supply of 25 volts was available to act as the raw supply for the zener diodes. For optimum zener diode performance, the raw supply should b e approximately two or more times the zener voltage. The combined zener voltage of the two diodes was approximately 17 volts, therefore, the raw supply voltage should have been 34 volts. Since this value of voltage was unavailable, the 25 volt supply was used.

In the precision laboratory, a test circuit was devised to obtain information on the following: drift, effects of ambient conditions, actual zener voltage, and possible circuit parameters. The circuit diagram is shown in Figure 3.1.

The actual measurements taken were of V_1 and V_2 (see Figure 3.1). These were taken by the use of the potentiometer, standard cell and galvanometer. The standard cell temperature correction chart provided was used in balancing the potentiometer. The data obtained is shown in TABLE I

and TABLE II.



$R_o = 100\Omega$

$R = \text{Variable}$

$R_1 = 16,150\Omega$

$R_2 = 850\Omega$

$Z_1 = Z_2 = 1N430A$

FIGURE 3.1

CIRCUIT OF ZENER DIODE TEST

The equipment used in the test of the zener diodes is listed below:

1.	Potentiometer	(H. Tinsley and Co., Ltd.)	Type - 3387B
			No. - 1111
2.	Cell	(Weston Standard Cell)	Type - 1268
			No. - 1297
3.	Galvanometer	(H. Tinsley and Co., Ltd.)	Type - 3387B
			No. - 1311
4.	Resistor R_o	(H. Tinsley and Co., Ltd.)	Type - 2541 L.F.I.
			No. - 88592
			Range- 0 to 100Ω
5.	Resistor R	(H. Tinsley and Co., Ltd.)	Type - 2541
			No. - 88479
			Range- 0 to 1110Ω
6.	Resistor R_1 and R_2	(H. Tinsley and Co.,Ltd.)	
	a) Voltage Divider		Type - 4281A
			No. -1083
			Range- 1000Ω
	b) Resistance Box		Type - 2541 L.F.I.
			No. - 88480
			Range- 0 to 11,111Ω
7.	Zener diodes ($Z_1 = Z_2$)	(Hoffman)	Type - 1N430A

TABLE I

DATA OF ZENER DIODE TEST

(See Figure 3.1 for Symbol Meanings)

Date	Time	V_1 volts	I_1 ma.	V_2 volts	I_2 ma.	I_z ma.	R ohms	V_z volts
Oct. 28/60	9:05 AM	1.0800	10.800	.84675	.9962	9.804	821.714	16.935
	9:10 AM	1.0900	10.900	.84685	.9963	9.904	813.116	16.937
	9:15 AM	1.1000	11.000	.84705	.9965	10.004	804.382	16.941
	9:20 AM	1.1100	11.100	.8472	.9967	10.103	796.000	16.944
	9:25 AM	1.1200	11.200	.8473	.9968	10.203	787.740	16.946

TABLE II

DATA TO DETERMINE DRIFT AND AMBIENT CONDITIONS

OF ZENER DIODES

Date	Time	V_1 volts	I_1 ma.	V_2 volts	I_2 ma.	I_z ma.	R ohms	V_z volts
Oct. 28/60	12:10 PM	1.1003	11.003	.84705	.9965	10.007	804.5	16.941
	2:30 PM	1.1006	11.006	.8469	.9963	10.010	804.5	16.938
	3:00 PM	1.1007	11.007	.8470	.9964	10.011	804.5	16.940
	3:30 PM	1.10065	11.0065	.8470	.9964	10.010	804.5	16.940
	4:00 PM	1.10075	11.0075	.8470	.9964	10.011	804.5	16.940
	4:30 PM	1.1011	11.011	.8470	.9964	10.015	804.5	16.940
	5:00 PM	1.1010	11.010	.8470	.9964	10.014	804.5	16.940
	6:30 PM	1.1013	11.013	.8470	.9964	10.017	804.5	16.940
	7:00 PM	1.10135	11.0135	.8470	.9964	10.017	804.5	16.940
	7:30 PM	1.1015	11.015	.8470	.9964	10.019	804.5	16.940
	11:30 PM	1.1023	11.023	.8470	.9964	10.027	804.5	16.940
Oct. 29/60	9:30 AM	1.1038	11.038	.8470	.9964	10.042	804.5	16.940
	10:40 AM	1.1038	11.038	.8470	.9964	10.042	804.5	16.940
	11:10 AM	1.10385	11.0385	.8470	.9964	10.042	804.5	16.940
	12:10 PM	1.1042	11.042	.8470	.9964	10.046	804.5	16.940
	1:05 PM	1.1043	11.043	.8470	.9964	10.047	804.5	16.940
	2:05 PM	1.1047	11.047	.8470	.9964	10.051	804.5	16.940
	3:10 PM	1.1048	11.048	.8470	.9964	10.052	804.5	16.940
	4:05 PM	1.1050	11.050	.8470	.9964	10.054	804.5	16.940
	5:00 PM	1.1053	11.053	.84705	.9965	10.057	804.5	16.941

TABLE II (continued)

Date	Time	V_1 volts	I_1 ma.	V_2 volts	I_2 ma.	I_z ma.	R ohms	V_z volts
Oct. 29/60	6:00 PM	1.1052	11.052	.8470	.9964	10.056	804.5	16.940
Oct. 30/60	9:30 AM	1.1071	11.071	.8470	.9964	10.075	804.5	16.940
	10:30 AM	1.1073	11.073	.8470	.9964	10.077	804.5	16.940
Oct. 31/60	10:15 AM	1.1103	11.103	.8471	.9965	10.107	804.5	16.942
	11:15 AM	1.1108	11.108	.8472	.9967	10.111	804.5	16.944
	12:15 PM	1.1108	11.108	.8472	.9967	10.111	804.5	16.944
	1:15 PM	1.1110	11.110	.8472	.9967	10.113	804.5	16.944
	2:15 PM	1.1109	11.109	.8472	.9967	10.112	804.5	16.944
	3:15 PM	1.1110	11.110	.8472	.9967	10.113	804.5	16.944
	5:15 PM	1.1114	11.114	.8472	.9967	10.117	804.5	16.944
	7:00 PM	1.118	11.118	.8472	.9967	10.121	804.5	16.944
Nov. 1/60	9:30 AM	1.1194	11.194	.1475	.9970	10.197	804.5	16.950
	12:30 AM	1.1190	11.190	.84725	.9968	10.193	804.5	16.945
	8:10 PM	1.1152	11.152	.8472	.9967	10.155	804.5	16.944
Nov. 2/60	10:00 AM	1.1172	11.172	.8472	.9967	10.175	804.5	16.944
	7:30 PM	1.1182	11.182	.8472	.9967	10.185	804.5	16.944
Nov. 3/60	11:15 AM	1.1195	11.195	.8472	.9967	10.198	804.5	16.944

Note: The fifth figure of the zener voltage (V_z) was read with an accuracy of ± 1.

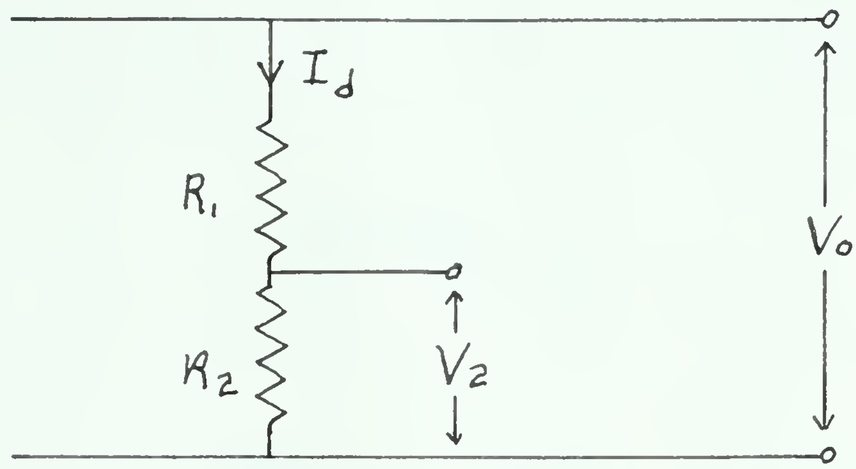
From the data in TABLE I it is seen that if the current through the diodes varied by ±1.82 per cent of 11.000 milliamperes, the zener voltage varied by +0.0295 per cent and -0.0354 per cent respectively. Also, if the diode current varied by ±0.91 per cent of 11.000 milliamperes, the zener voltage varied by +0.0177 per cent and -0.0236 per cent, respectively. The given relations do not indicate that the diodes are unsymmetrical. The fifth figure of the given zener voltages may be in error by ±1, as was noted on the data, and this amount caused the results to look unsymmetrical. The values given above are used as an indication of the zener regulation.

From the data in TABLE II it is seen that there was no variation in zener voltage over a period of about a day. However, there was a slight upward drift of the zener voltage over a period of approximately a week. This could have resulted from diode-aging and was not critical to the overall design. It is seen that over a short period of operation, the period over which the supply would probably be used, the zener diodes would provide adequate stability. Also, from the results it was seen that the stability of the raw supply was adequate. The zener voltages were read with an accuracy of 0.006 per cent.

The circuit diagram of the actual device used is

shown in Figure 3.3

Figure 3.2 shows the circuit of the Voltage Divider.



R_1 = 900,000Ω (wire wound)

R_2 = 15,000Ω plus 1000Ω trimmer (wire wound)

V_o = 1000 V.

FIGURE 3.2

VOLTAGE DIVIDER

The Voltage Divider was designed to provide a proportion of the output voltage equal to the zener voltage, which in this case was very nearly 17 volts. Also, the Divider was to draw approximately one milliampere from the output. Because of the above conditions and the resistors available, the Voltage Divider was constructed with the resistance values R_1 equal to 900,000 ohms and R_2 equal to 15,000 ohms plus a 1000 ohm trimmer. Resistance R_2 was made

up of one 15,000 ohm resistor and one 1000 ohm trimmer. The resistors used in the Voltage Divider were wire wound to obtain temperature stability. The trimmer was also wire wound. With the trimmer set at zero ohms the conditions are as follows (see Figure 3.2 for symbol meanings):

$$V_2 = \frac{15,000}{15,000 + 900,000} V_o$$

$$V_2 = 16.40 \text{ V}$$

$$I_d = \frac{1000}{R_1 - R_2} = \frac{1000}{915,000} = 1.094 \text{ ma.}$$

With the trimmer set at the maximum of 1000 ohms the conditions are,

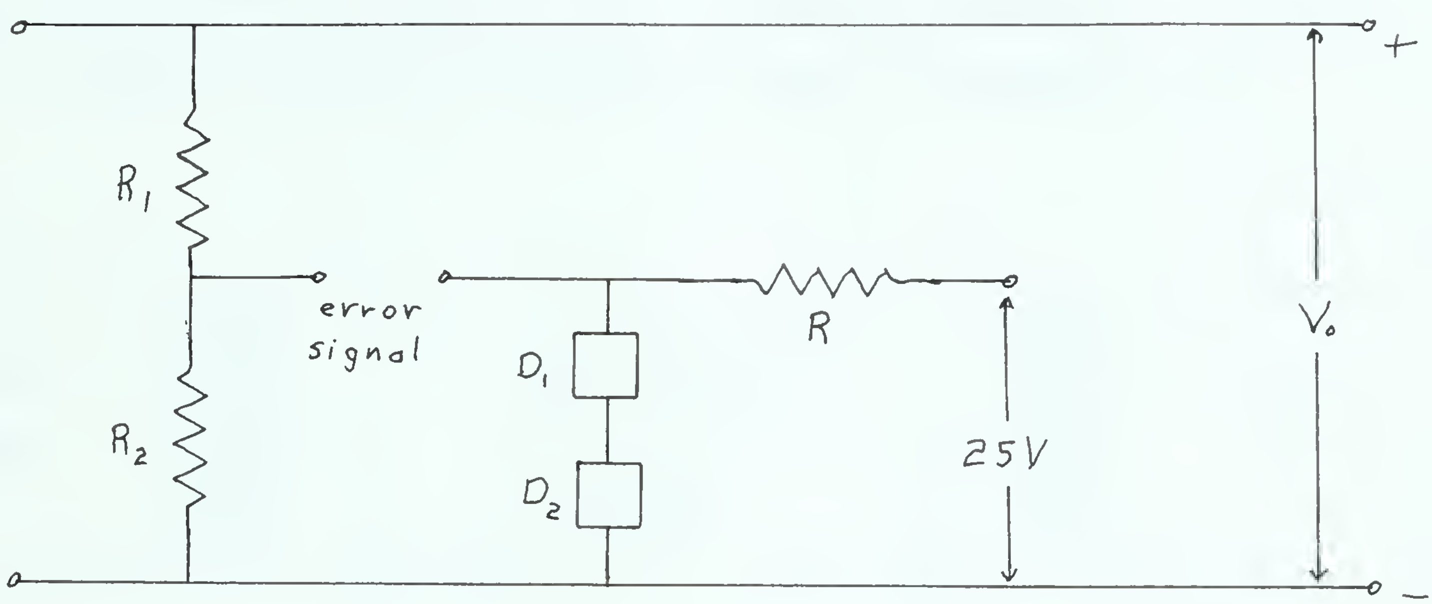
$$V_2 = \frac{16,000}{16,000 + 900,000} V_o$$

$$V_2 = 17.48 \text{ V}$$

$$I_d = \frac{1000}{916,000} = 1.093 \text{ ma.}$$

A voltage variation from 16.40 to 17.48 volts can be obtained by the use of the trimmer. This voltage range includes the desired value, which is about 17 volts. The amount of current drawn is also satisfactory.

Figure 3.3 shows the scheme used to compare the Reference Voltage to the voltage from the Voltage Divider.



$R = 800\Omega$ (wire wound)

$R_1 = 900{,}000\Omega$ (wire wound)

$R_2 = 15{,}000\Omega$ plus 1000Ω trimmer (wire wound)

$D_1 = D_2 = $ 1N430A

FIGURE 3.3

REFERENCE VOLTAGE AND VOLTAGE DIVIDER

3.3 RING MODULATOR

Being a relatively simple device, no design calculations were required for the Ring Modulator. Figure 3.4 shows a circuit diagram of the arrangement used.

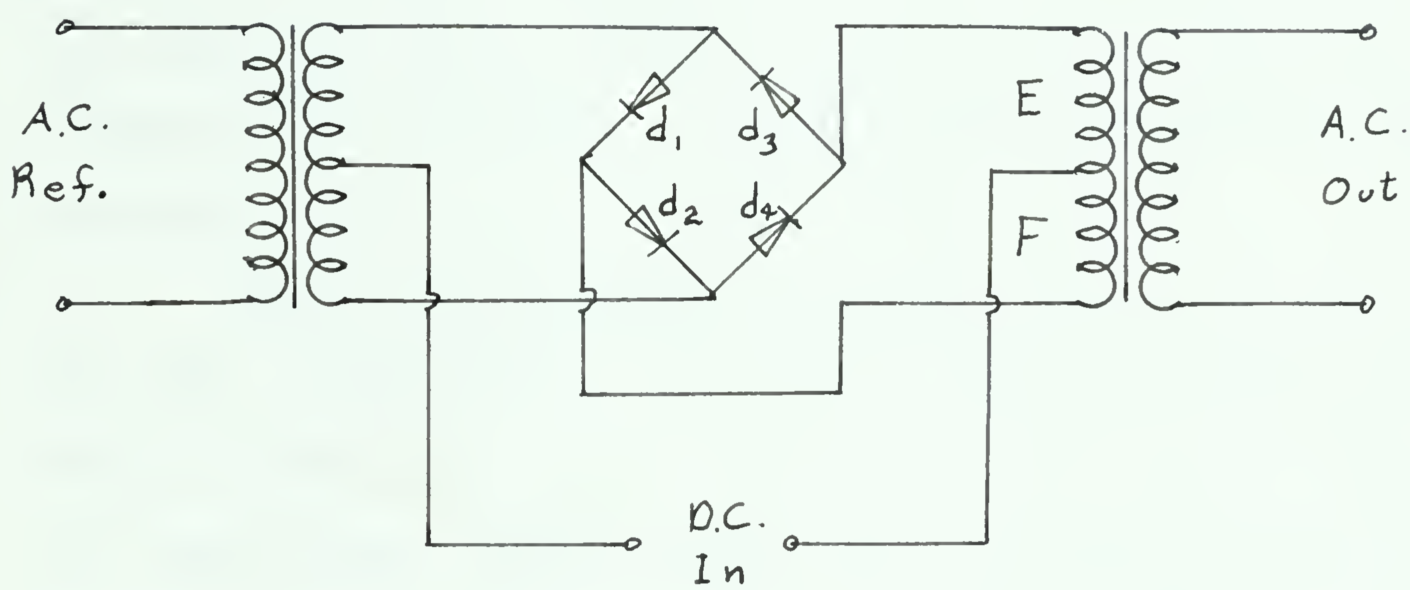


FIGURE 3.4

RING MODULATOR

To aid the description of operation of the Ring Modulator the diodes will be considered as switches. During one half of the reference cycle when diodes d_1 and d_2 are closed and diodes d_3 and d_4 are open, D.C. signal passes through the F half of the output transformer. During the other half of the reference cycle, the diode switching is reversed and

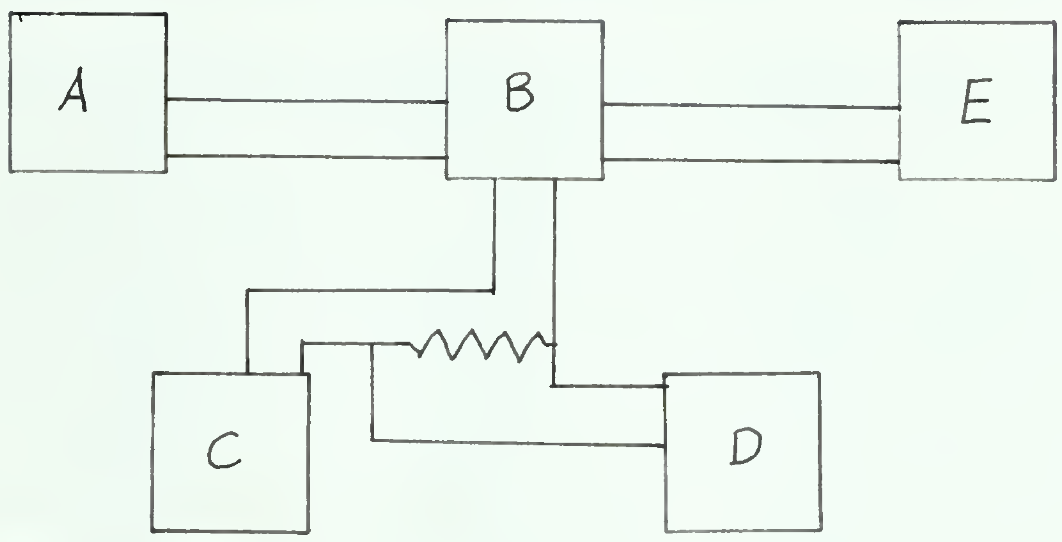
the D.C. signal passes through the E half of the output transformer. The resulting signal at the output of the modulator is A.C. with an amplitude that is dependent on the D.C. input.

In the ideal case one would have a square wave acting as the reference signal. This would make the diodes act more like ideal switches, consequently, diode matching would be less critical. A square wave generator was not used because it requires special transformers.

Two center-tapped transformers were hand wound on 3B, D25, 16 Ferroxcube pot cores. The transformers were wound bifilar in order to achieve a better match between the two halves. Also, shields were included between the primary and secondary windings to reduce capacitive coupling between the two windings. The shield consisted of a single-layer winding, of which one terminal was grounded. In actual operation it was noted that these shields provided a great improvement on the zero signal conditions. The diodes which were available and proved satisfactory were the silicon 1N459 diodes. Since a number of these diodes were available it was possible to match two pairs of these diodes by the use of the Transistor Curve-Tracer, to further reduce the output signal at zero input. The chosen A.C. reference was at a frequency

of 10,000 cycles per second.

The Modulator was constructed as shown in Figure 3.4. The output impedance of the Ring Modulator was determined experimentally. A block diagram of the scheme used is shown in Figure 3.5



A = Oscillator

B = Ring Modulator

C = Simulated error

D = Voltmeter

E = Variable load

FIGURE 3.5

SCHEME TO DETERMINE OUTPUT IMPEDANCE
OF THE MODULATOR

TABLE III presents the data obtained from the experiment. The conditions set for this experiment are listed below:

1. A.C. reference input = 1 V. (peak to peak).
2. A.C. reference frequency = 10,000 cps.
3. Simulated error input = 0.1 μa D.C.

TABLE III

DATA TO OBTAIN MODULATOR OUTPUT IMPEDANCE

V_{out}	Load Imped.	Power Out
P. to P. mv.	ohms	Watts
9.5	50 K	2.26×10^{-10}
13.5	70 K	3.25×10^{-10}
17.5	90 K	4.24×10^{-10}
20.5	100 K	5.24×10^{-10}
24.0	120 K	6.00×10^{-10}

The results are shown graphically in Figure 3.6.

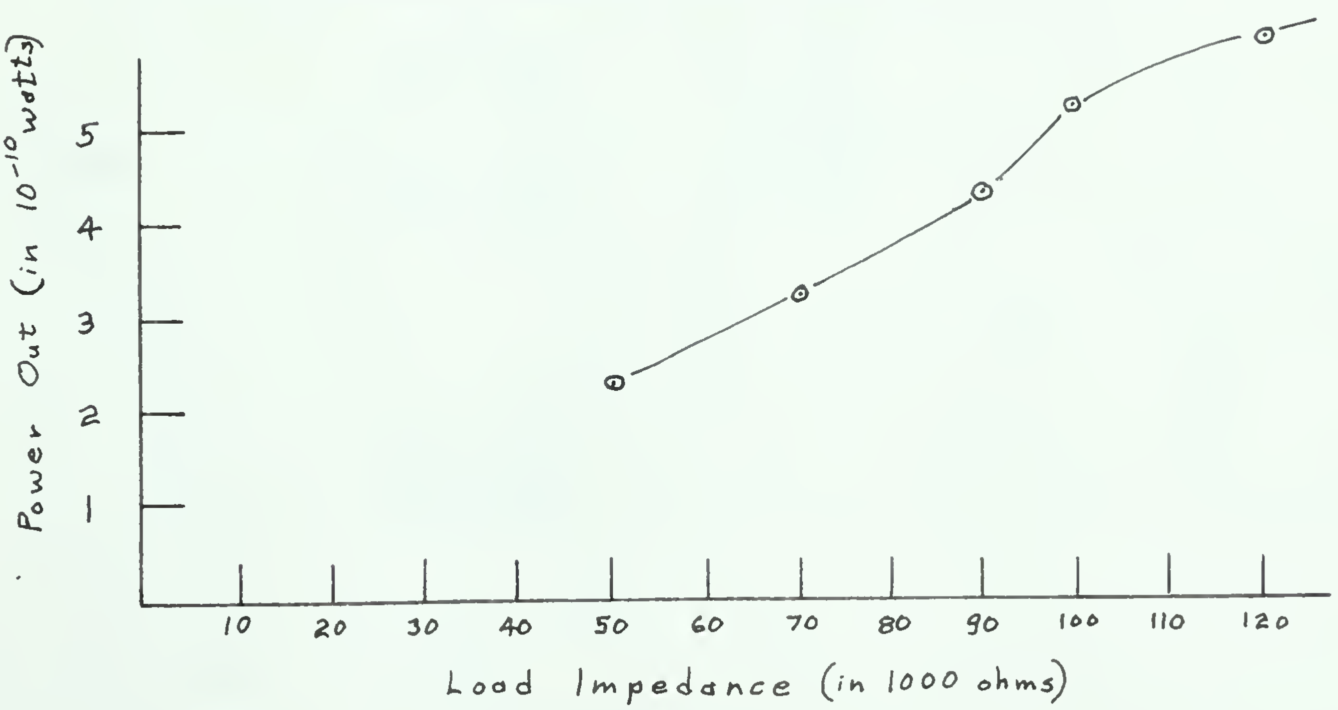


FIGURE 3.6

IMPEDANCE CHARACTERISTIC OF THE RING MODULATOR

To avoid extremely high impedances, the load impedance of the Modulator was chosen to be 80,000 ohms.

The output of the Modulator terminated into the Error Amplifier and because of the relatively high output impedance of the Modulator, the first stage of the Amplifier was an emitter-follower. To avoid stray pick-up due to high impedance termination between component boxes, the emitter-follower stage was built in the same component box as the Modulator, with a metallic shield around it. A circuit diagram of the actual Modulator and emitter-follower stage is shown in Figure 3.7. The design of the emitter-follower stage is shown in the design of the Error Amplifier. The impedance of the output of the emitter-follower was calculated as 2000 ohms.

Q = 2N43A
R_1 = 2000Ω
R_2 = 5600Ω
R_3 = 18,000Ω
C_1 = 0.25 μfd.
$d_1 = d_2 = d_3 = d_4$ = 1N459
T_{11} = 600 turns
T_{12} = 300 turns
T_{21} = 300 turns
T_{22} = 600 turns

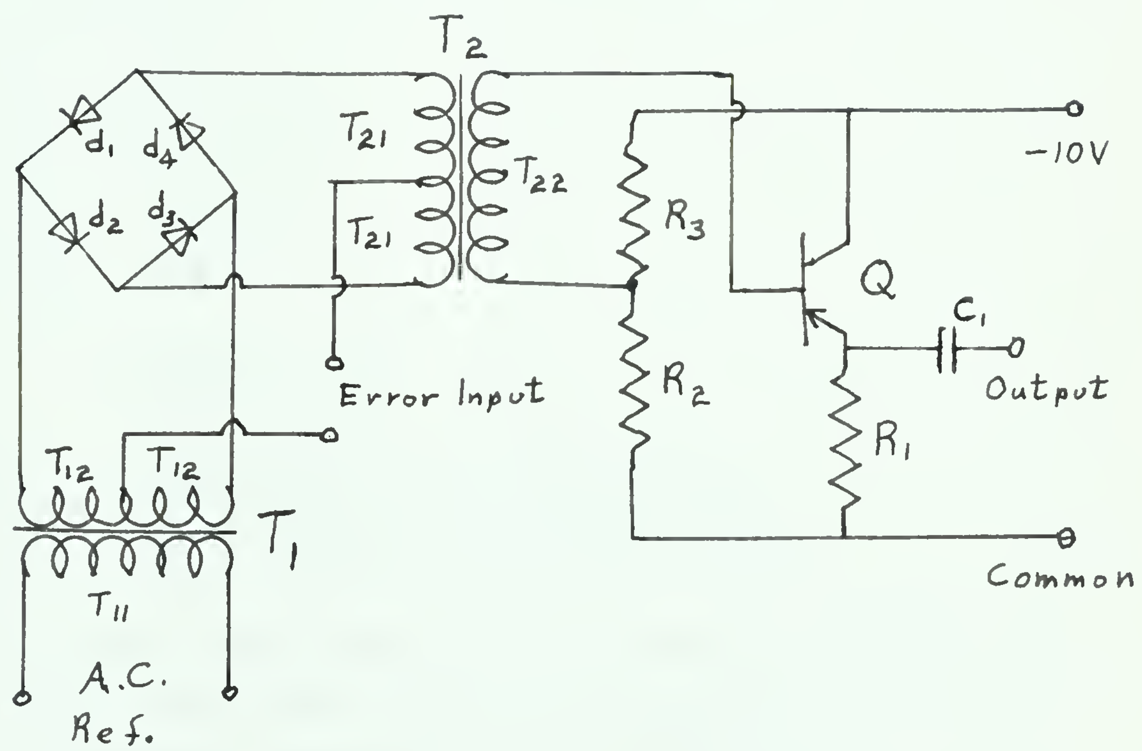
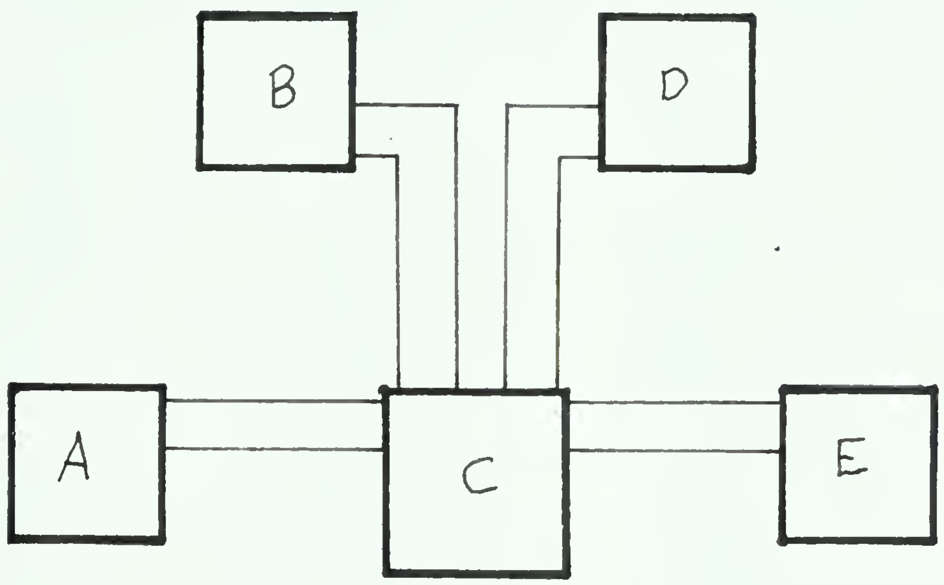


FIGURE 3.7

ACTUAL MODULATOR AND EMITTER-FOLLOWER STAGE

To determine the sensitivity of the Ring Modulator and the emitter-follower stage another test was conducted. A block diagram describing the scheme used is shown in Figure 3.8.



A = Oscillator
B = Simulated error
C = Ring modulator
D = Supply volts
E = 2000Ω load

FIGURE 3.8

SCHEME USED TO DETERMINE THE SENSITIVITY OF THE MODULATOR AND EMITTER-FOLLOWER STAGE

The data obtained from the test is shown in TABLE IV. The conditions set for this test are listed below:

1. A.C. reference frequency = 10,000 c.p.s.
2. A.C. reference voltage = 0.25 V.
3. Supply voltage = 10 V.

TABLE IV

DATA USED TO DETERMINE SENSITIVITY OF THE MODULATOR AND EMITTER-FOLLOWER

D.C. Error in μa	V_{out} (P-P) across 2KΩ	Power-Out (Watts)
.00	0.4 mv	1.0×10^{-11}
.001	1.0 "	6.27 x "
.002	1.4 "	12.25 x "
.003	1.7 "	18.0 x "
.005	1.9 "	22.6 x "
.01	2.2 "	30.3 x "
.02	2.6 "	42.1 x "
.03	3.0 "	56.2 x "
.04	3.4 "	72.2 x "
.05	3.9 "	95.2 x "
.06	4.4 "	121.3 x "
.07	4.8 "	144.5 x "
.08	5.2 "	169.0 x "
.09	5.6 "	196.0 x "
.10	6.0 "	225.0 x "
- 0.0	0.4 mv	1.0×10^{-11}
- .001	.8 "	3.97 x "
- .002	.9 "	5.05 x "
- .003	1.1 "	7.56 x "
- .005	1.4 "	12.25 x "
- .01	1.8 "	20.2 x "
- .02	2.2 "	30.3 x "
- .03	2.6 "	42.1 x "
- .04	3.0 "	56.2 x "
- .05	3.6 "	81.0 x "
- .06	4.0 "	100.0 x "
- .07	4.4 "	121.3 x "
- .08	5.0 "	156.7 x "
- .09	5.4 "	182.5 x "
- .10	6.0 "	225.0 x "

The data of Table IV is shown graphically in Figure 3.9.

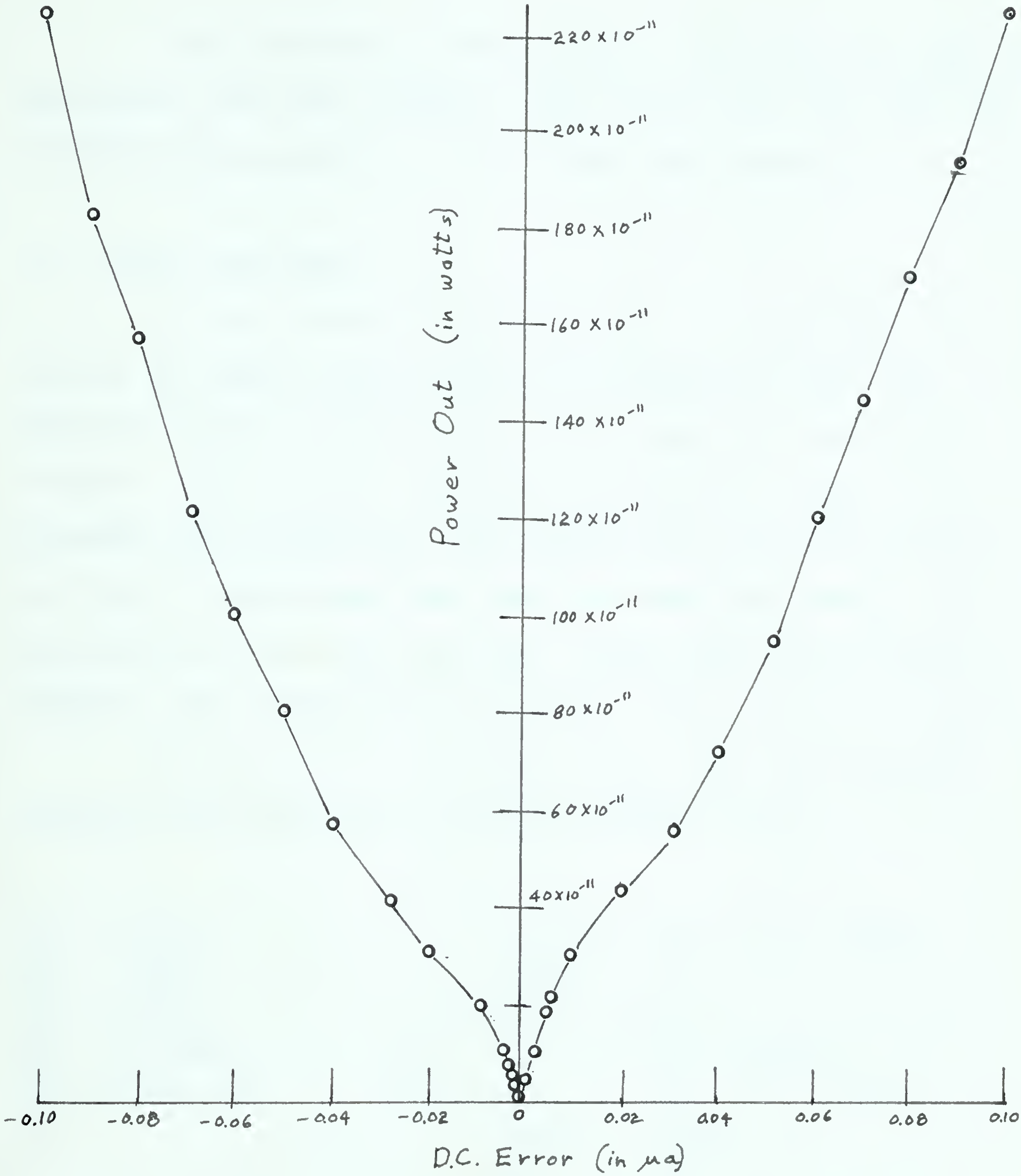


FIGURE 3.9

GRAPH SHOWING THE SENSITIVITY OF RING MODULATOR
AND EMITTER-FOLLOWER

From the graph in Figure 3.9 it was concluded that the device was sensitive enough. Also, it was seen that the sensitivity was greatest at the smaller input signals.

3.4 ERROR AMPLIFIER

At this stage of the design, it was impossible to obtain the desired gain of the Error Amplifier since the characteristics of some of the other components were still unknown. As a result, the Amplifier was built with an adjustable gain, therefore, the needed gain could be obtained as long as the maximum gain available was high enough. As indicated earlier, the Amplifier design consisted of three stages: two emitter-follower stages, and a push-pull stage.

The circuit of the first stage, which was included with the Ring Modulator is shown in Figure 3.10.

Q = 2N43A
R_1 = 2000Ω
R_2 = 5600Ω
R_3 = 18,000Ω
C_1 = 0.25 μfd.

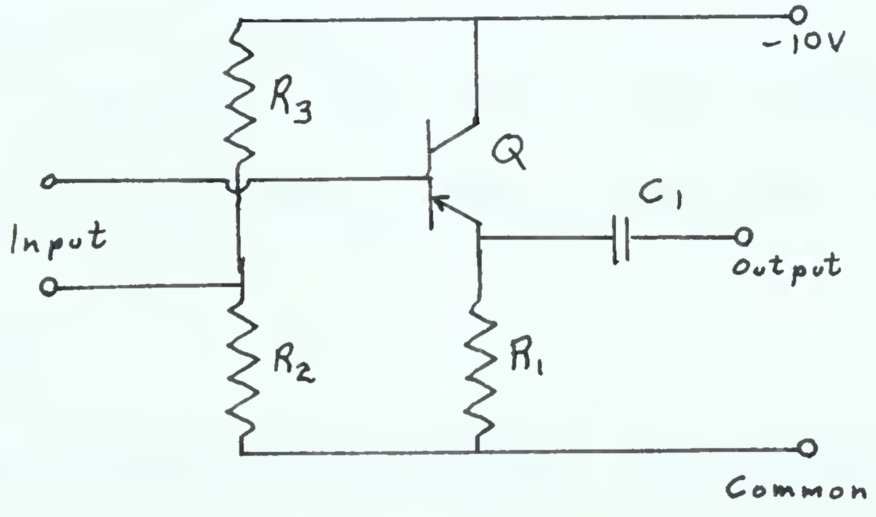


FIGURE 3.10

FIRST EMITTER-FOLLOWER STAGE

Transistors 2N43A were available and were used for this amplifier. From the characteristic curves, (see Appendix B), the following operating conditions were chosen:

$I_c = 1$ ma.

$I_b = 30\ \mu$a.

$V_{cc} = -10$ V.

A value for R_1 of 2000 ohms was considered. The current through R_1 was,

$1 - 0.03 = 0.97$ ma.

Therefore, the voltage drop across it was,

$0.97 \times 2 = 1.94$ V.

As a result the collector voltage, neglecting the base to emitter voltage, was approximately,

$10 - 1.94 = 8.06$ V.

From these values the transistor dissipation was approximately,

$8.06 \times 1 = 8.06$ mw.

Since the specified maximum dissipation is 150 milliwatts the above condition was satisfactory. Also, 1.94 volts provided sufficient range for the A.C. voltage swing. Therefore, the value was,

$\underline{R_1 = 2000\ \Omega}$

The calculations of R_2 and R_3 follow: (see equations in Appendix A)

Equation G: $$s = \frac{1}{1 - \alpha + \alpha\left(\frac{R_1}{(R_1 + R_B)}\right)}$$

Chose, $s = 3$

Since R_1 is known and α is given in Appendix B,

$$\underline{R_B = 4330\ \Omega}$$

Equation E:

$$I_B = \frac{V_{CC}}{R_3} - \frac{I_C R_1 + V_{EB}}{R_B}$$

By estimating V_{EB} to be 0.2 volts,

$$R_3 = 18{,}500\ \Omega$$

Value used,

$$\underline{R_3 = 18{,}000\ \Omega}$$

Equation B: $$R_B = \frac{R_3 R_2}{R_3 + R_2}$$

Since R_B and R_3 are known,

$$R_2 = 5700\ \Omega$$

Value used,

$$\underline{R_2 = 5600\ \Omega}$$

The calculation of the output stage follows:

$$Z_o = (1 - \alpha)\ Z_{in}\ *$$

*<u>The Radio Amateur's Handbook</u>, (American Radio Relay League, Inc., U.S.A., 1960), p. 82.

Since Z_{in} is 80,000 ohms (see Modulator design) and $(1 - \alpha)$ is 0.025 (see Appendix B),

$$\underline{Z_o = 2000\ \Omega}$$

The value of C_1 was calculated by making its reactance equal to 5 per cent of Z_o. Therefore,

$$X_{C1} = 0.05 \times 2000 = 100\ \Omega$$

$$= \frac{1}{2\pi f C_1} \qquad \text{where } f = 10{,}000 \text{ c.p.s.}$$

$$C_1 = 0.16 \times 10^{-6} \text{ fd.}$$

Value used,

$$\underline{C_1 = 0.25\ \mu\text{fd.}}$$

The design of the second stage was identical to the first stage, except the stability factor S was chosen as 3.5 instead of 3. The resulting component values calculated and used are shown in Figure 3.11.

Q_d = 2N43A
R_{d1} = 2000Ω pot.
R_{d2} = 7500Ω
R_{d3} = 15,000Ω
C_d = 1 μfd.
T_{11} = 150 turns
T_{12} = 900 turns

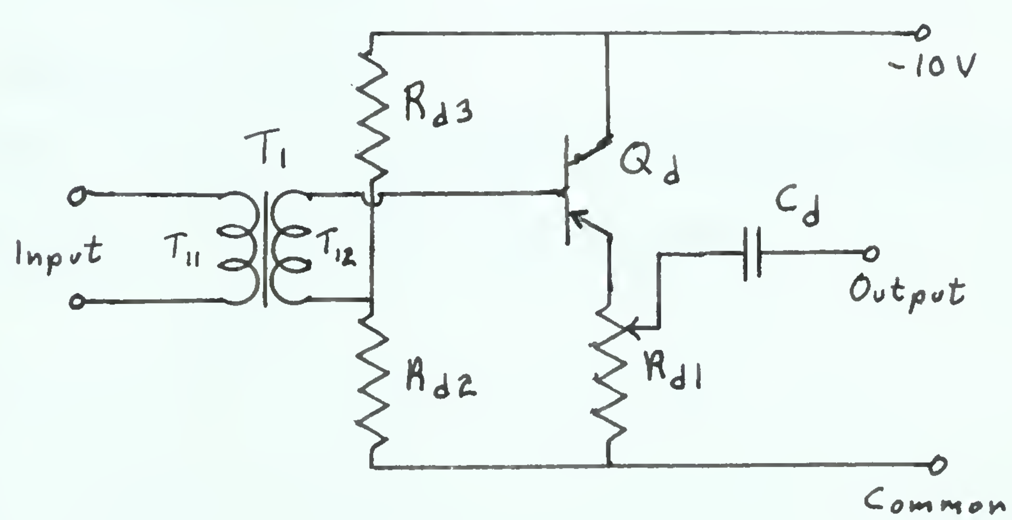


FIGURE 3.11

SECOND EMITTER-FOLLOWER STAGE

Instead of a 2000 ohm resistor for R_{d1}, a 2000 ohm potentiometer was used so that the gain of the amplifier could be varied.

The turns ratio for transformer T_1 was calculated as follows:

$$Z_o \text{ of first stage} = 2000\ \Omega$$

$$Z_{in} \text{ of second stage} = 80,000\ \Omega$$

Therefore, $T_{12} = \frac{80,000}{2000} = 6.32.$

Value used,

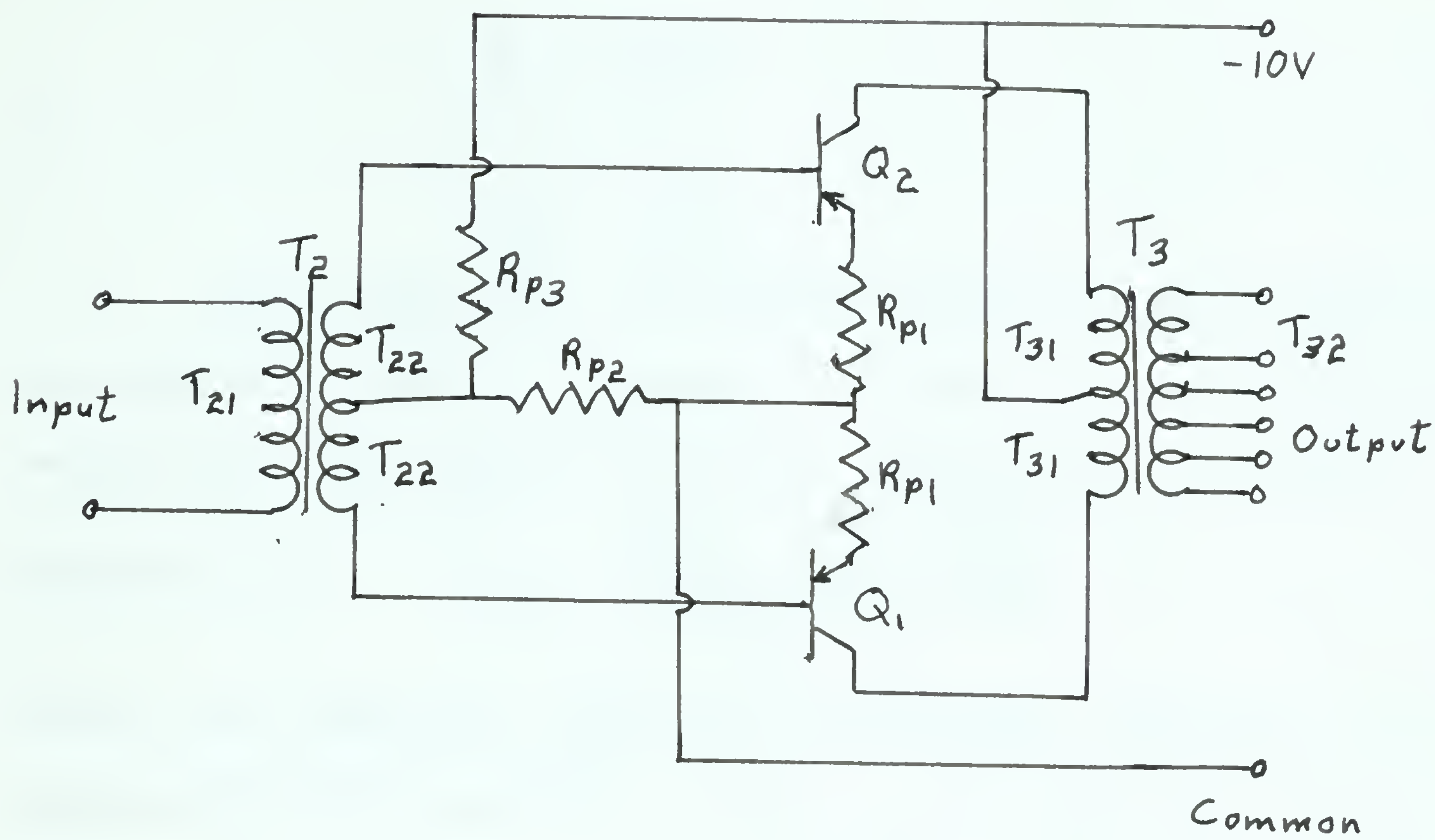
$$\frac{T_{12}}{T_{11}} = 6$$

A Ferroxcube pot core 3B, D25, 16 was used. Using number 36 wire, about 1200 turns could be wound on the bobbin. As a result the turns used were,

$$T_{12} = 900 \text{ turns}$$

$$T_{11} = 150 \text{ turns}$$

The push-pull stage was designed for class A operation and its circuit diagram is shown in Figure 3.12.



$Q_1 = Q_2 = 2N43A$
$R_{p1} = 250\Omega.$
$R_{p2} = 1800\Omega.$
$R_{p3} = 5600\Omega.$
$T_{21} = 600$ turns
$T_{22} = 300$ turns

$T_{31} = 400$ turns
$T_{32} = 67$ turns
= 80 turns
= 100 turns
= 134 turns
= 200 turns
(the secondary (T_{32}) was tapped)

FIGURE 3.12

PUSH-PULL STAGE OF ERROR AMPLIFIER

From the characteristic curves (see Appendix B), the following operating conditions were chosen:

$$I_c = 8 \text{ ma}.$$

$$I_b = 120\ \mu a.$$

$$V_c = -8\ V.$$

$$V_{cc} = -\ 10\ V.$$

The values for the biasing resistors are calculated by equations given in Appendix A. From the chosen operating conditions, the value of R_{p1} was calculated as follows:

Equation F: $$R_{p1} = \frac{(V_{CC} - V_{CE} - I_C R_L)}{I_C + I_B}$$

Since this amplifier stage was transformer coupled, its D.C. load resistance was neglected. Also, I_B was neglected. Therefore,

$$R_{p1} = 250\ \Omega.$$

Value used,

$$\underline{R_{p1} = 250\ \Omega.}$$

The value of R_B was calculated next.

Equation G: $$S = \frac{1}{1 - \alpha + \alpha\left(\frac{R_1}{(R_1 + R_B)}\right)}$$

Chose, $S = 3.5$

Where $\alpha \equiv 0.975$ (see Appendix B)

$$R_1 = 250\ \Omega.$$

Therefore,

$$\underline{R_B = 1310\ \Omega.}$$

The value of R_{p3} was calculated next.

Equation E: $$I_B = \frac{V_{CC}}{R_{p3}} - \frac{I_C R_1 + V_{CE}}{R_B}$$

Where V_{CE} 0.2 V.

Since the only unknown is R_{p3} its value was found to be,

$$R_{p3} = 5560\ \Omega.$$

Value used,

$$R_{p3} = 5600\ \Omega.$$

The calculations for R_{p2} were done next.

Equation B: $$R_B = \frac{R_{p2}R_{p3}}{R_{p2} + R_{p3}}$$

Knowing R_{p3} and R_B

$$R_{p2} = 1800\ \Omega.$$

Value used,

$$\underline{R_{p2} = 1800\ \Omega.}$$

The turns ratio for transformer T_2 was found next. Typical values for the input and output impedances for a 2N43A transistor were obtained.*

Typical Z_{in} of 2N43A = 600 Ω.

Typical Z_o of 2N43A = 30,000 Ω.

Therefore, the input impedance of the push-pull stage was chosen as 600 ohms. The output impedance of the previous stage was found to be 2000 ohms. Therefore,

*Marrow H.E., Transistor Engineering Reference Handbook, (John F. Rider Publisher, Inc., N.Y., 1956) p, II-35

$$\frac{T_{21}}{T_{22}} = \frac{2000}{600} = 1.83$$

Value used,

$$\frac{T_{21}}{T_{22}} = 2.$$

The actual turns used were,

$$\underline{T_{21} = 600 \text{ turns}}$$

$$\underline{T_{22} = 300 \text{ turns}}$$

The output impedance of the push-pull stage was found next.

$$Z_o = \frac{1}{h_{22}\left(1 - \dfrac{h_{11}S}{h_{11} + Z_g}\right)} \quad **$$

Where $S = \dfrac{h_{12}h_{21}}{h_{11}h_{22}}$

From the hybrid parameters given in Appendix B.

$$S = \frac{2.23 \times 10^{-4} \times 43.5}{1246 \times 22.3 \times 10^{-6}}$$

$$S = .335$$

Therefore, the calculated output impedance was,

$$\underline{Z_o = 59{,}000\ \Omega.}$$

** Lowry H.R. and Associates, General Electric Transistor Manuel, (Canadian General Electric Co., Ltd., Toronto, 1960), p. 31.

The load impedance to which the output of this amplifier terminated was unknown, therefore, the turns ratio was calculated for a number of estimated impedances and the output of the transformer was tapped at these values.

The estimated load impedances and the corresponding turns ratios are listed:

1. $Z_L = 1640\ \Omega,\ \dfrac{T_{31}}{T_{32}} = \sqrt{\dfrac{59{,}000}{1640}} = 6$

2. $Z_L = 2360\ \Omega,\ \dfrac{T_{31}}{T_{32}} = 5$

3. $Z_L = 3680\ \Omega,\ \dfrac{T_{31}}{T_{32}} = 4$

4. $Z_L = 6600\ \Omega,\ \dfrac{T_{31}}{T_{32}} = 3$

5. $Z_L = 14{,}800\Omega,\ \dfrac{T_{31}}{T_{32}} = 2$

The resulting turns were:

$T_{31} = 400$ turns

$T_{32} = 67$ turns

$= 80$ turns

$= 100$ turns

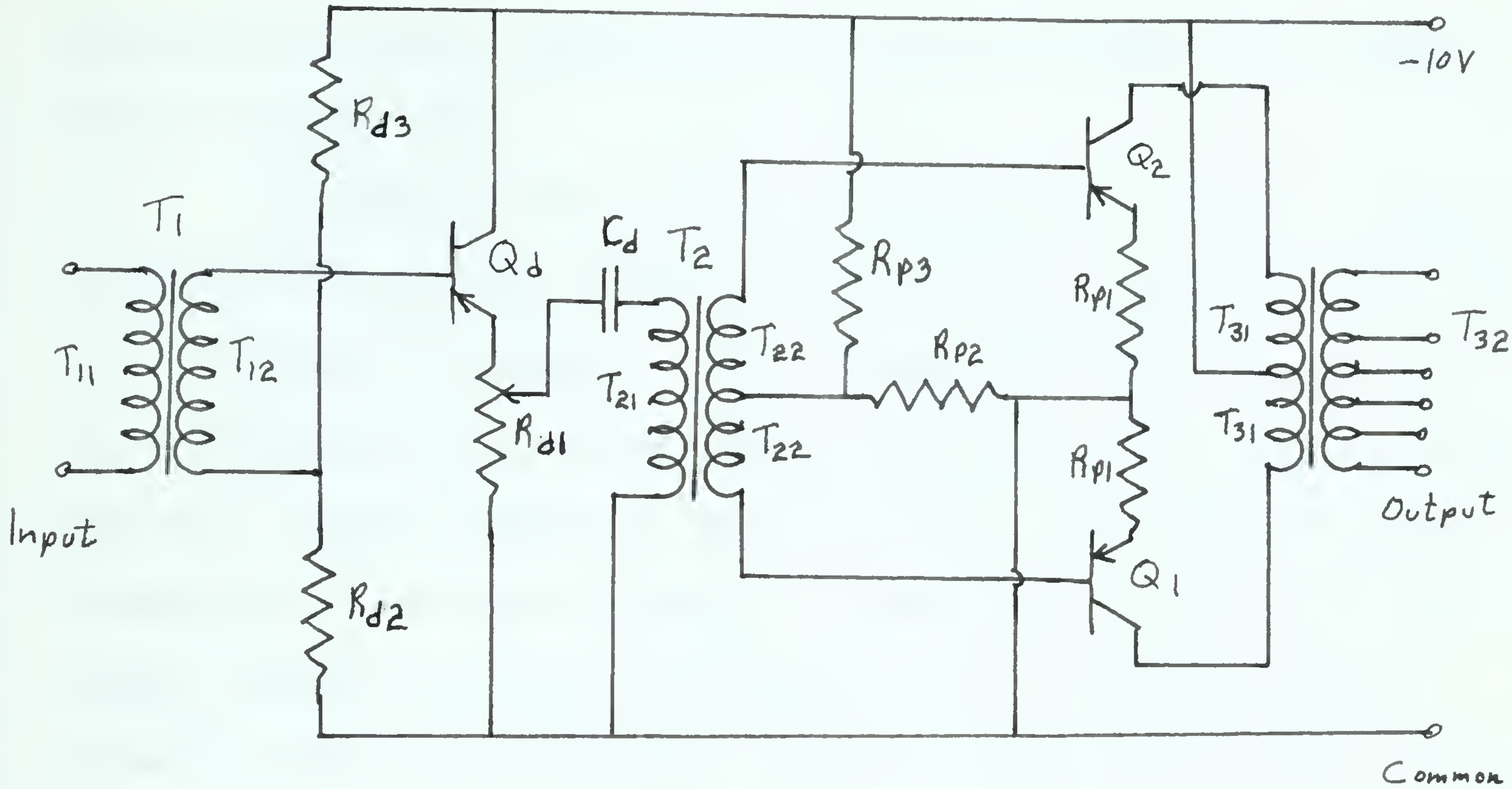
$= 134$ turns

$= 200$ turns.

Experimentally it was found that the best value was,

$\underline{T_{32} = 200 \text{ turns}}$

Figure 3.13 shows the circuit of the last two stages of the Error Amplifier as they were built.



$Q_d = Q_1 = Q_2$ = 2N43A	T_{12} = 900 turns	T_{31} = 400 turns
R_{d1} = 2000 Ω pot.	R_{p1} = 250 Ω	T_{32} = 67 turns
R_{d2} = 7500 Ω	R_{p2} = 1800 Ω	= 80 turns
R_{d3} = 15,000 Ω	R_{p3} = 5600 Ω	= 100 turns
C_d = 1 μfd.	T_{21} = 600 turns	= 134 turns
T_{11} = 150 turns	T_{22} = 300 turns	= 200 turns

FIGURE 3.13

ERROR AMPLIFIER

3.5 RING DEMODULATOR

The construction of the Ring Demodulator was similar to that of the Ring Modulator. However, there were small resistors included in series with the diodes to limit the A.C.

reference current, since the A.C. reference was at a higher level in this case.

In addition to the Ring Demodulator circuit, a stage of amplification and a phase shifter were included before the A.C. reference terminals. The reason for this was that the A.C. reference of the Demodulator had to be at a higher level and at a certain phase to the Modulator. The amplification stage was an emitter-follower, of which the calculated component values are shown in Figure 3.14. A resistor and a capacitor were used across a center-tapped transformer, as a phase-shifter. The circuit values were obtained experimentally. The Ring Demodulator, phase shifter and amplifier were built as one unit, of which the circuit is shown in Figure 3.14.

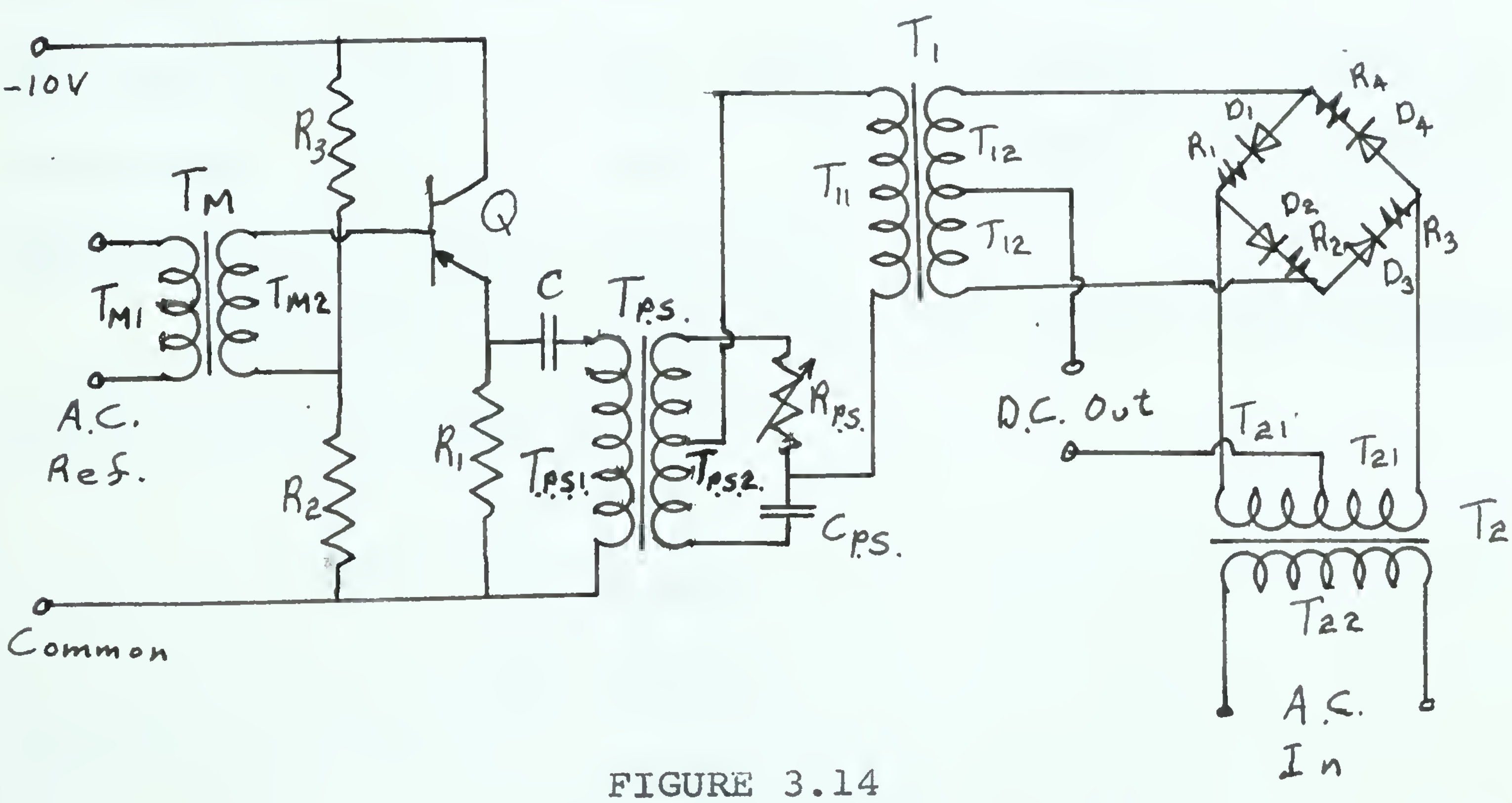


FIGURE 3.14

ACTUAL RING DEMODULATOR CIRCUIT

T_{M1} = 250 turns	Q = 2N43A
T_{M2} = 250 turns	R_1 = 330 Ω
T_{PS1} = 250 turns	R_2 = 1500 Ω
$T_{PS\ 2}$ = 500 turns	R_3 = 3900 Ω
T_{11} = 400 turns	C = 1 μfd.
T_{12} = 400 turns	R_{PS} = 5000 Ω (variable)
T_{21} = 300 turns	C_{PS} = 0.022 μfd.
T_{22} = 300 turns	$D_1 = D_2 = D_3 = D_4$ = 1N459A
	$R_1 = R_2 = R_3 = R_4 = 33\ \Omega$.

FIGURE 3.14

LIST OF COMPONENTS

3.6 REFERENCE OSCILLATOR

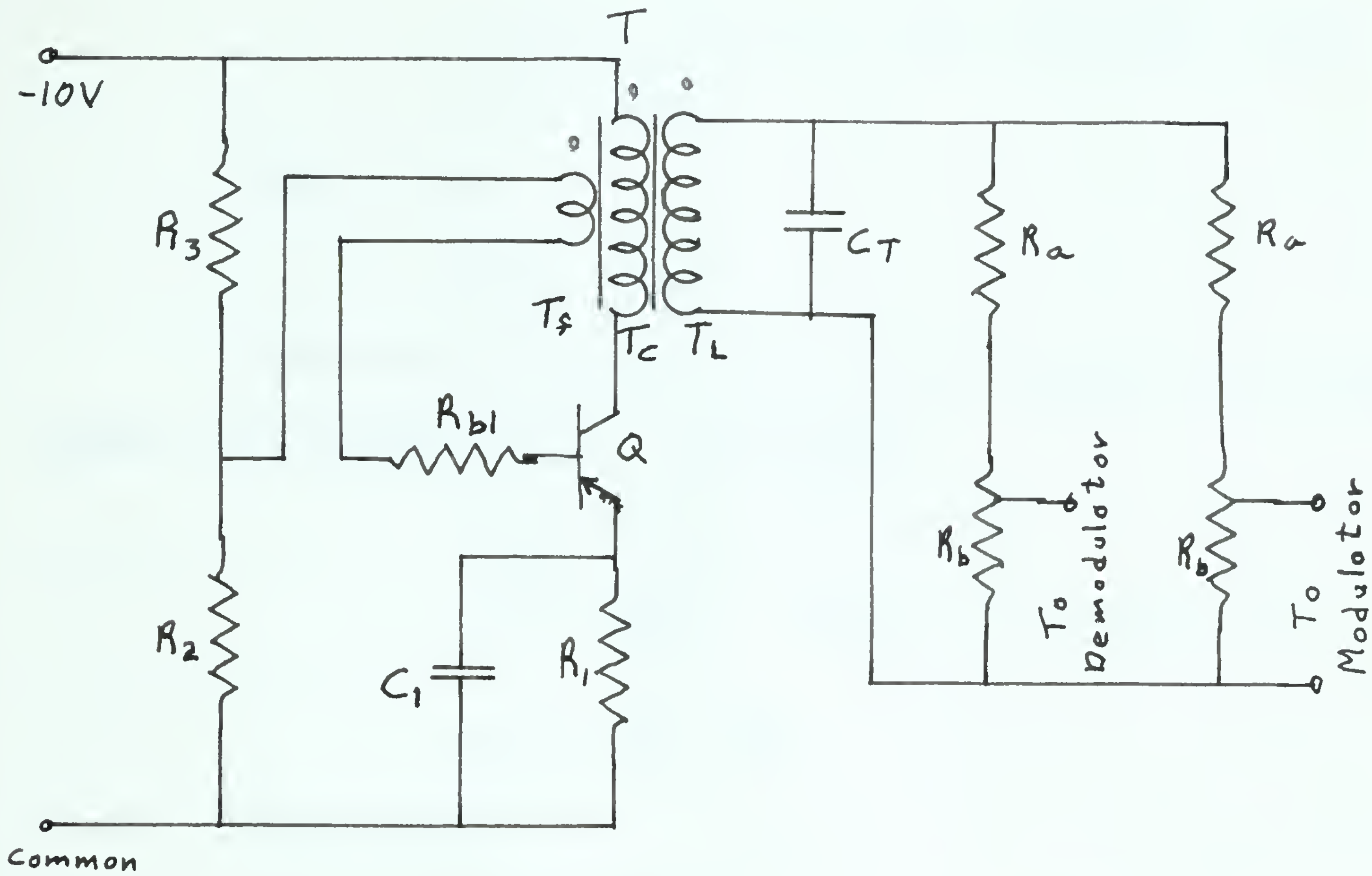
In order to obtain maximum temperature stability a silicon transistor was used for the Reference Oscillator. The Oscillator was designed by designing an amplifier first and then applying the proper amount of feedback. To aid the discussion Figure 3.15 shows a circuit diagram of the Reference Oscillator, as it was constructed.

The chosen operating conditions follow (see characteristic curves in Appendix B):

$$I_c = 4 \text{ ma.}$$

$$I_b = 90\ \mu\text{a.}$$

$$V_{cc} = -10 \text{ V.}$$



Q = 2N118	T_f = 8 turns
R_1 = 500 Ω	T_C = 200 turns
R_2 = 1600 Ω	T_L = 1000 turns
R_3 = 5600 Ω	C_T = 154 μμfd.
R_{b1} = 1000 Ω	R_a = 100,000 Ω
C_1 = 1 μfd.	R_b = 10,000 Ω pot.

FIGURE 3.15

REFERENCE OSCILLATOR

The value of resistor R_1 was found by making the power dissipated in it equal 20 per cent of the total D.C. power. Therefore, by assuming that $I_e \approx I_c$,

$$(4 \times 10^{-3})^2 R_1 = .20 \times 4 \times 10^{-3} \times 10$$

$$R_1 = 500 \ \Omega.$$

Value used,

$$\underline{R_1 = 500 \ \Omega}$$

The value of C_1 was obtained by making its reactance equal to 5 percent of R_1. Therefore,

$$X_{C1} = 0.05 \times 500 = 25 \ \Omega$$

$$25 = \frac{1}{2\pi C_1 f} \qquad \text{where } f = 10{,}000 \text{ c.p.s.}$$

$$C_1 = .637 \ \mu\text{fd.}$$

Value used,

$$\underline{C_1 = 1 \ \mu\text{fd.}}$$

The values of R_2 and R_3 were calculated by equations in Appendix A.

Equation G: $$S = \frac{1}{1 - \alpha + \alpha\left(\frac{R_1}{(R_1 + R_B)}\right)}$$

Chose, $S = 3.5$

Where, $\alpha = 0.975$ (see Appendix B)

$R_1 = 500 \ \Omega.$

Therefore,

$$\underline{R_B = 1250 \ \Omega.}$$

The value of resistor R_3 was obtained next.

Equation E: $$I_B = \frac{V_{CC}}{R_3} - \frac{I_C R_1 + V_{BE}}{R_B}$$

Where $V_{BE} \approx 0.22$

Therefore,

$$R_3 = 5380\ \Omega.$$

Value used,

$$\underline{R_3 = 5600\ \Omega.}$$

The calculation of R_2 follows:

Equation B: $$R_B = \frac{R_2 R_3}{R_2 + R_3} = 1250\ \Omega.$$

Therefore,

$$R_2 = 1610\ \Omega.$$

Value used,

$$\underline{R_2 = 1600\ \Omega.}$$

The next step was to estimate the impedance into which the Reference Oscillator would terminate. To do this an oscillator was connected across the terminals of the Ring Modulator, since the Modulator was part of the load, and the approximate operating conditions were established. The results obtained were that the A.C. reference voltage across the Modulator was 0.18 volts and the current was 2.25 microamperes. Therefore, the impedance of the modulator reference was calculated as 80,000 ohms. Since the Demodulator was not built yet, the combined impedance of the Demodulator and the Modulator was estimated to be 40,000 ohms. This value was used as the load impedance of the Oscillator.

Having the load impedance and the common-emitter hybrid parameters for the 2N118 (see Appendix B), the input impedance was calculated.

$$Z_{in} = h_{11} - \frac{h_{21}h_{12}Z_L}{1 + h_{22}Z_L} \quad *$$

Where,

$$h_{11} = 4500\ \Omega$$
$$h_{21} = 36.7$$
$$h_{12} = 5 \times 10^{-4}$$
$$h_{22} = 2 \times 10^{-5}\ \text{mhos.}$$
$$\underline{Z_{in} = 4090\ \Omega.}$$

Next, the current gain was calculated.

$$A_i = \frac{h_{21}}{1 + h_{22}Z_L} \quad **$$

$$\underline{A_i = 20.4}$$

Since no other conditions were known, it was assumed that the output and input impedances are resistive. Knowing the two impedances and the current gain, the power gain was calculated.

*Lowry H.R. and Associates, General Electric Transistor Manual, (Canadian General Electric Co. Ltd., Toronto, 1960),p.31

** Ibid.

$$G = (A_i)^2 \frac{Z_o}{Z_{in}} = (20.4)^2 \frac{40,000}{4090}$$

$\underline{G = 4,070}$.

Next, the transformer turns were established. Since the Oscillator was to be terminated to a relatively high resistance and also since it is easier to tune a coil with a larger number of turns, the load coil was chosen to have 1000 turns. The coil in the collector circuit was chosen to have 200 turns.

To calculate the number of turns in the feedback coil, the procedure used is given below:

$$\frac{\text{Power to load}}{\text{Power to feedback}} \approx 4070.$$

Therefore,

$$\frac{V_L^2}{Z_L} \Big/ \frac{V_f^2}{Z_{in}} = 4070.$$

Again, Z_L and Z_{in} are assumed resistive,

Or, $\dfrac{N_L^2}{Z_L} \times \dfrac{Z_{in}}{N_f^2} = 4070$

$$N_f^2 = \frac{N_L^2 \times Z_{in}}{4070 \times Z_L}$$

Since, $Z_{in} = 4090\ \Omega.$

$Z_L = 40,000\ \Omega.$

$N_L = 1000$ turns

Then, $N_f^2 = 25.1$

$$N_f = 5.01 \text{ turns}$$

Allowing a safety factor for the feedback, the value for the feedback turns was chosen as,

$$\underline{N_f = 8 \text{ turns.}}$$

This transformer was wound on a Ferroxcube pot core, 3B, D25, 16. The inductance of the load coil was measured as 1.65 henries.

The calculations for the tuning capacitor C_T follow:

$$f_c = \frac{1}{2\pi\ LC} = 10{,}000 \text{ c.p.s.}$$

$$C_T = \frac{1}{(2\pi f)^2\ L}$$

$$\underline{C_T = 154\ \mu\mu\text{fd.}}$$

A trimmer condenser was used to provide the necessary capacity.

With the biasing affected only be resistors R_1, R_2, and R_3, the output of the Oscillator was somewhat distorted. Experimentally it was found that by inserting a 1000 ohm resistor in series with the transistor base, the base bias was decreased and the waveshape was improved. Therefore,

$$R_B = 1000\ \Omega.$$

Across the output coil of the oscillator, two resistances were connected in parallel. Each resistance consisted

of a 100,000 ohm resistor and a 10,000 ohm potentiometer. The Modulator and Demodulator reference terminals were connected to the potentiometer taps. This arrangement provided an output of about 4 volts (peak to peak) which was adequate. Also, variations in the Oscillator load due to variations in the Modulator or Demodulator would have negligible effects.

3.7 D.C MIXER

Three D.C. voltages had to be summed to form the supply voltage of the Master Oscillator. Since the Demodulator load had to be small the voltages were put in series at the input of an emitter-follower, D.C. amplifier. Figure 3.16 shows the circuit as it was constructed.

Q = 2N327A(Si.)

R_b = 1000 Ω

R_e = 15,000

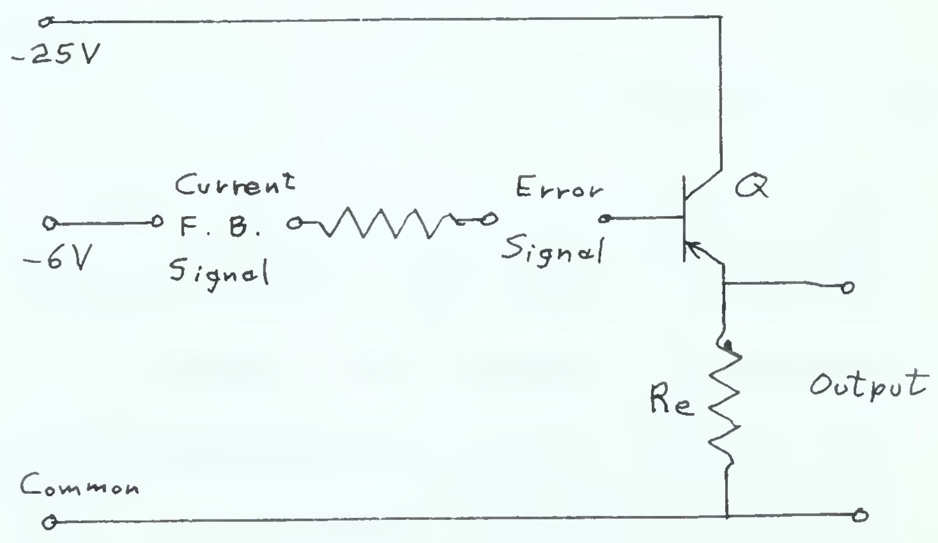


FIGURE 3.16

D.C. MIXER

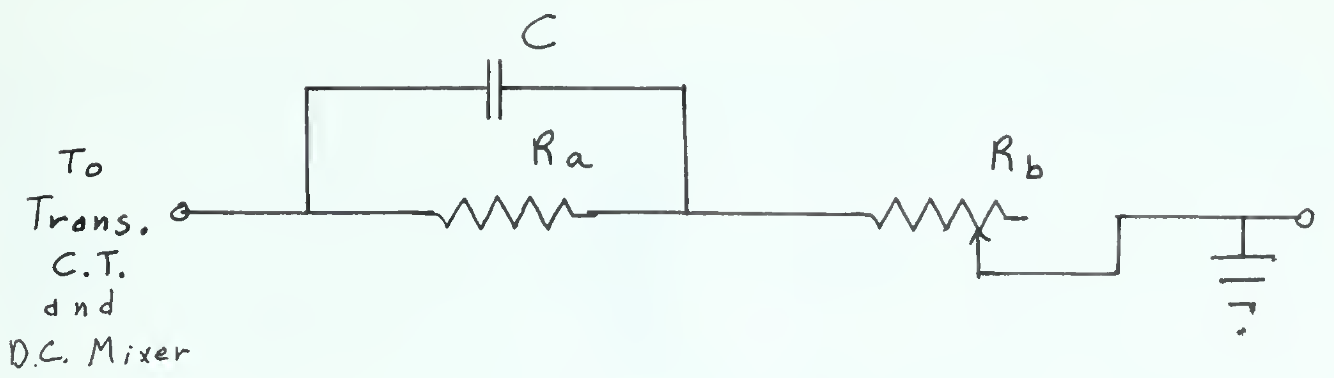
The circuit was obtained experimentally and to check the design, the collector current was measured at maximum output and was found to be 5 milliamperes. At the extreme when the current feedback voltage and the error voltage were zero, the collector voltage was 19 volts. The resulting transistor dissipation was,

$$19 \times 5 = 95 \text{ mw.}$$

This included a safety factor, because when the current feedback voltage and the error voltage were zero the output current was less than 5 milliamperes. The manufacturer's specifications state a maximum collector current of 100 milliamperes and a maximum power dissipation of 350 milliwatts. Therefore, operating conditions of the circuit were satisfactory.

3.8 CURRENT FEEDBACK DEVICE

To feed back a voltage proportional to the load current, a resistor was inserted at the center-tap of the Power Transformer. If only a resistor were used, there was a blocking-oscillator effect so that the output oscillated. To minimize this oscillation, part of the resistance was A.C.-shorted by a relatively large capacitor. A variable resistor was used so that the amount of feedback could be adjusted. The values used were determined experimentally. The circuit of the actual Current Feedback Device is shown in Figure 3.17.



$C = 100\ \mu fd.$

$R_a = 250\ \Omega$ (wire wound)

$R_b = 1000\ \Omega$ (wire wound pot.)

FIGURE 3.17

CURRENT FEEDBACK DEVICE

3.9 MASTER OSCILLATOR

The design of the Master Oscillator does not follow the scheme in Appendix A. This was one of the first components built and the type of design was not established at the time. The circuit diagram is shown in Figure 3.18.

A 2N43A transistor was used. The chosen operating conditions are given below (see Appendix B):

$I_c = 2.5$ ma.

$I_b = 45\ \mu a.$

$V_{cc} = -10$ V.

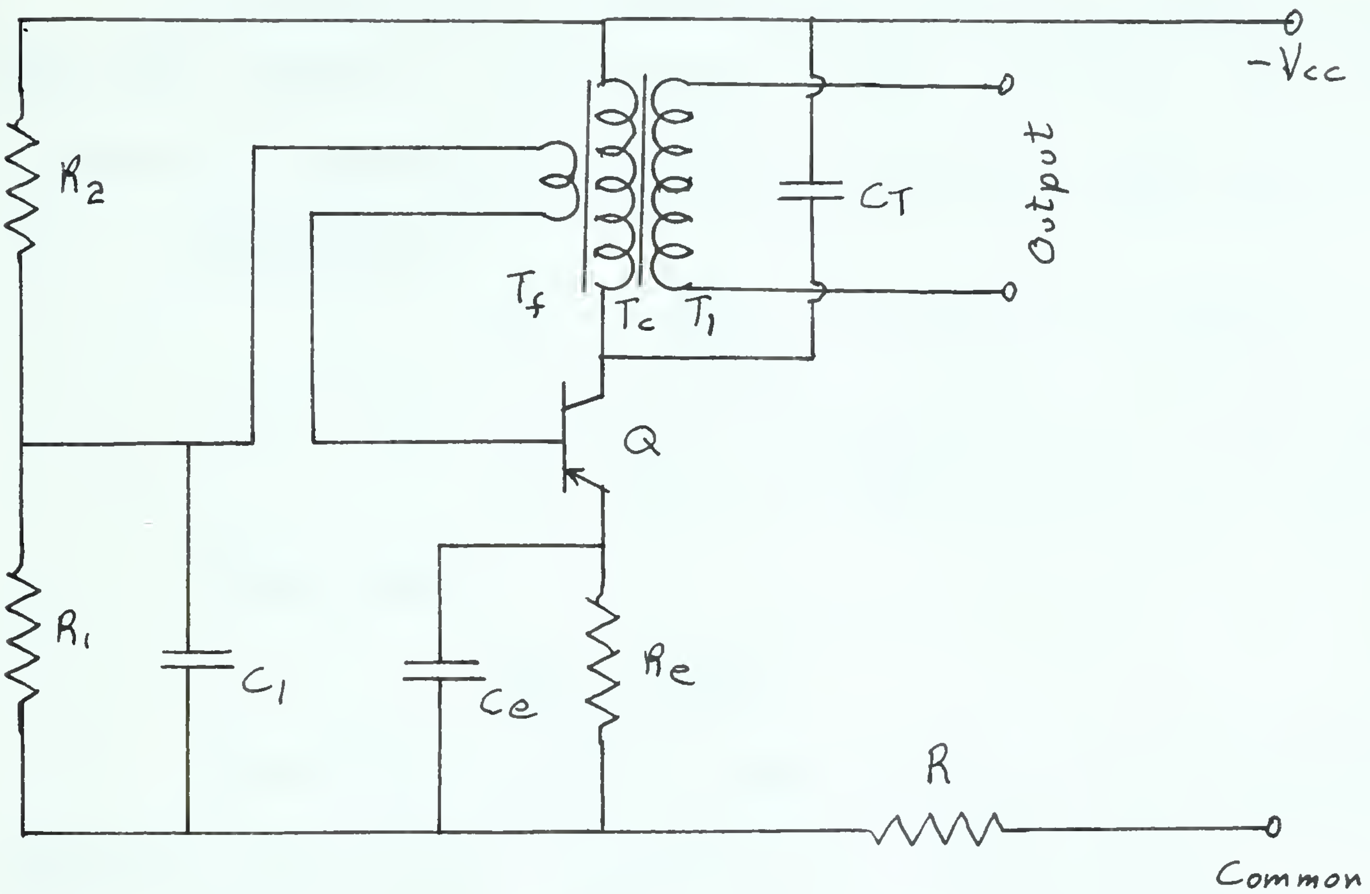


FIGURE 3.18

MASTER OSCILLATOR

Resistor R_e was calculated by making the power dissipated in it equal to 10 per cent of the D.C. power used. Therefore,

$$I_e^2 R_e = .10 \times 2.5 \times 10 \times 10^{-3}$$

Since $\quad I_e \approx I_c$

$$R_e = 400\ \Omega.$$

Value used,

$$\underline{R_e = 330\ \Omega.}$$

To prevent A.C. degeneration a by-pass capacitor C_e was used across R_e. The value of this capacitor was chosen so that its capacitive reactance was about 5 per cent of R_e.

Therefore,

$$X_{ce} = \frac{1}{2\pi f C_e} = 0.05 \times 330, \text{ where } f = 4000 \text{ c.p.s.}$$

$$C_e = \frac{1}{2\pi \times 4 \times 10^3 \times 16.5} = 2.41 \ \mu\text{fd.}$$

Value used,

$$\underline{C_e = 4 \ \mu\text{fd.}}$$

Values for R_1 and R_2 were obtained by calculating a value for R_2 and then making R_1 ten times larger.

The voltage across R_2 was approximately equal to the supply voltage minus the drop across R_e (neglecting V_{EB}).

Therefore,

$$V_{R2} = 10 - 2.5 \times 10^{-3} \times 330$$

$$V_{R2} = 9.2 \text{ V.}$$

Allowing 45 μa. through R_2, to provide the biasing, the value of R_2 was,

$$R_2 = \frac{9.2}{45 \times 10^{-6}} = 205{,}000 \ \Omega.$$

Value used,

$$\underline{R_2 = 200{,}000 \ \Omega.}$$

The value of R_1 follows:

$$R_1 = 10 \times R_2$$

Value used,

$$\underline{R_1 = 2M\,\Omega.}$$

By inserting resistor R_1 into the circuit the chosen base current was decreased slightly. However, since R_1 was very large as compared to the emitter resistor R_e, the decrease in the bias current could be neglected.

The value of C_1 was chosen rather arbitrarily by making it comparable to C_e. Therefore,

$$\underline{C_1 = 2\ \mu\text{fd.}}$$

No calculations were done to obtain the turns ratio for the transformer. The number of turns of the load coil and the collector coil were both chosen to be 500 turns. The number of turns of the feedback coil was obtained experimentally and was found to be 4 turns. Having the transformer built, it was found that the inductance of the collector coil, which was the one to be tuned, was 0.303 henries. The value of capacity needed to tune this oscillator to 4000 cycles per second was calculated as,

$$f_c = \frac{1}{2\pi\ LC_T}$$

$$C_T = \frac{1}{(2\pi f)^2 L} = \frac{1}{(2\pi\ 4000)^2 \times .303}$$

$$\underline{C_T = 0.00523\ \mu\text{fd.}}$$

A trimmer condenser in parallel with a fixed condenser were used to obtain the required capacity.

Having this oscillator built it was found that its output was somewhat distorted. It was found that if a 33,000 ohm resistor was put in series with the supply voltage, the bias was reduced and some degenerative feedback was provided. The resulting waveshape of the output was good.

It was required that the output of this oscillator varied linearly with the supply voltage. To determine whether this was so, the supply voltage was varied from zero to 20 volts and the output voltage was measured across a 70,000 ohm load. The data obtained is shown in TABLE V. The results are also shown graphically in Figure 3.19. From these results it is seen that the output to input voltage relation is linear.

TABLE V

DATA TO OBTAIN LINEARITY OF MASTER OSCILLATOR

V_{cc} D.C. Volts	V_{Load} V_{p-p}	V_{cc} D.C. Volts	V_{Load} V_{p-p}
0.0	0.0	13.0	25.10
1.0	1.60	14.0	27.20
2.0	3.70	15.0	29.30
3.0	5.60	16.0	31.40
4.0	7.60	17.0	33.50
5.0	9.60	18.0	35.70
6.0	11.50	19.0	37.90
7.0	13.40	20.0	40.10
8.0	15.30		
9.0	17.20		
10.0	19.10		
11.0	21.00		
12.0	23.00		

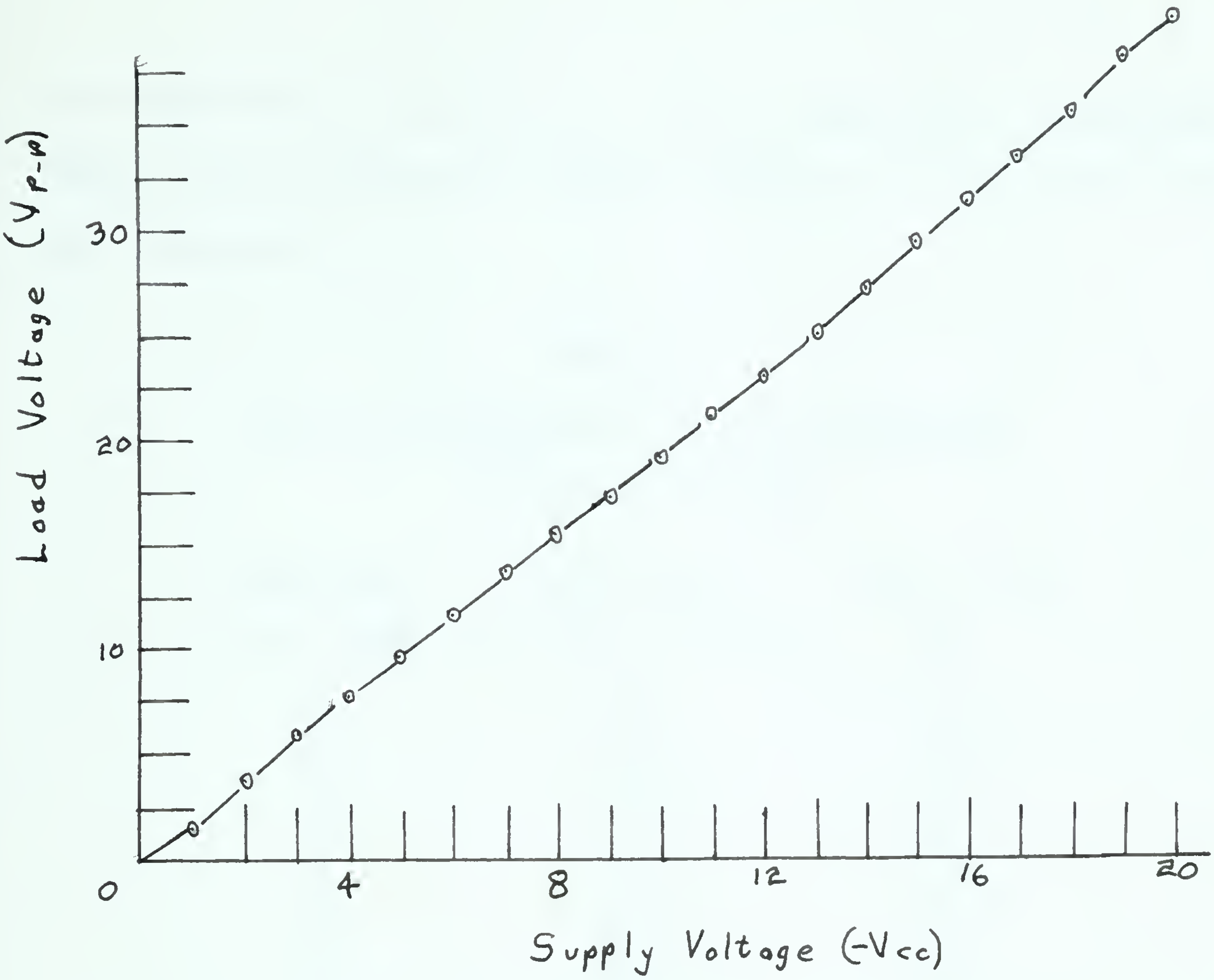


FIGURE 3.19

STATIC RESPONSE OF MASTER OSCILLATOR

The load impedance into which the oscillator should terminate was determined experimentally. To do this the supply voltage was held constant at -15 volts and the load impedance was varied. By measuring the voltage across the load impedance, the power output was calculated. The data obtained is shown in TABLE VI and represented graphically in Figure 3.20. From the graph in Figure 3.20 it was decided

to terminate the oscillator into an impedance of 100,000 ohms since this is the most stable region as far as load variations are concerned.

TABLE VI

DATA TO DETERMINE THE LOAD IMPEDANCE OF THE OSCILLATOR

Load Res. in K ohms	Load Volts. P. to P. Volts.	Power Out mw.
10	19.0	4.52
20	23.0	3.30
30	25.5	2.71
40	27.0	2.28
50	28.0	1.96
60	28.5	1.70
70	29.0	1.50
80	29.5	1.36
90	30.0	1.25
99	30.5	1.18

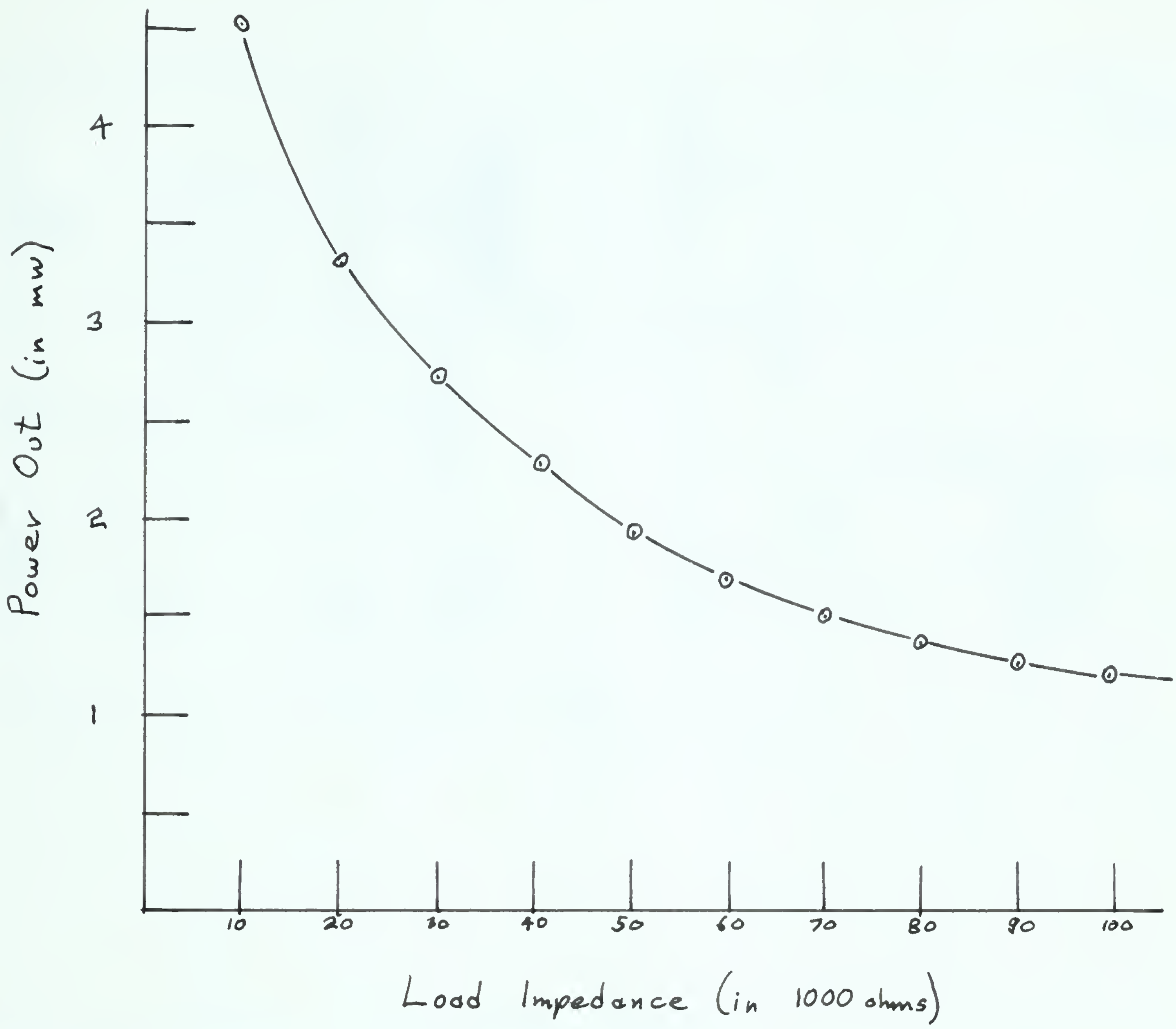
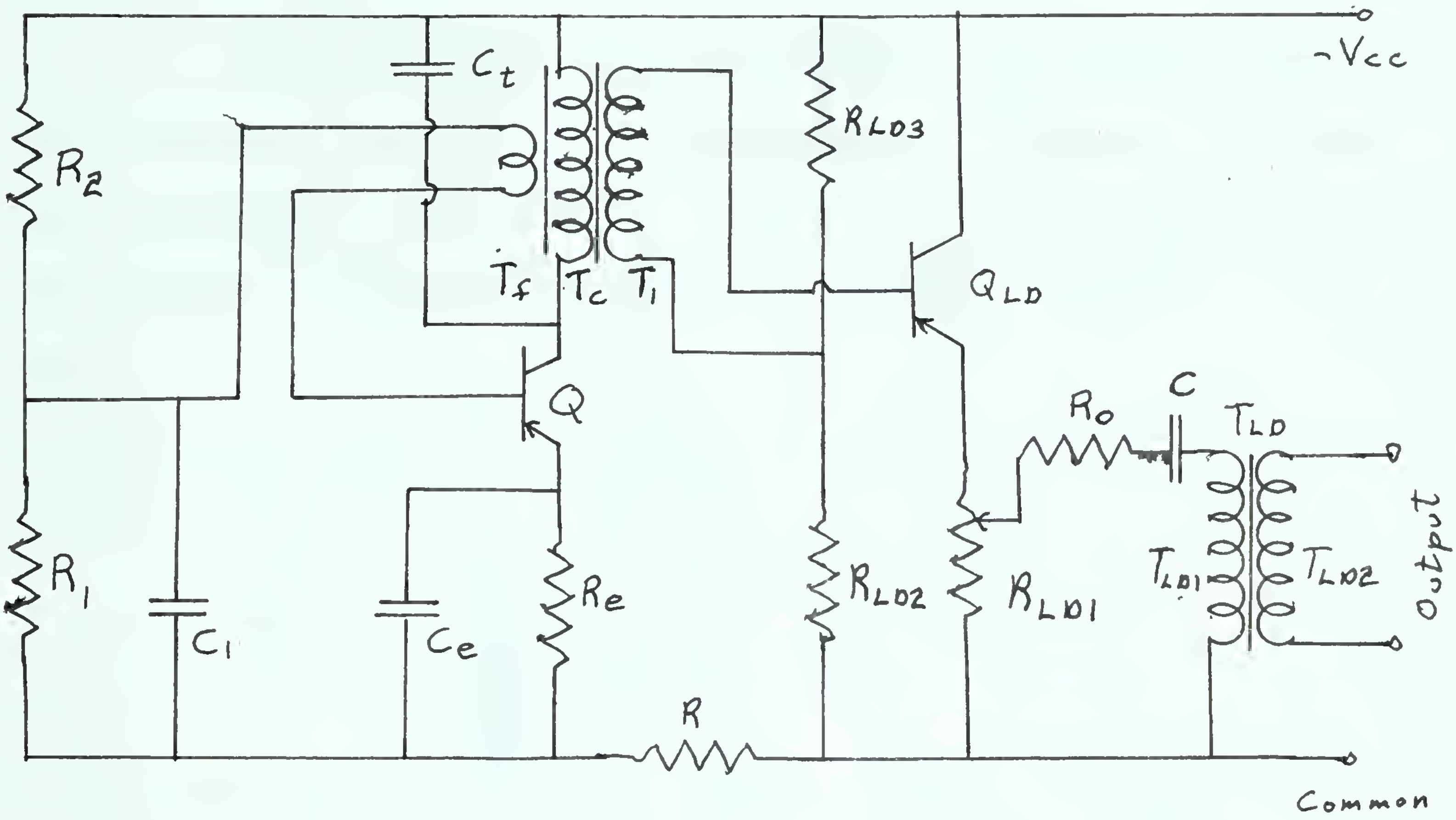


FIGURE 3.20

GRAPH OF THE EFFECT OF VARIATION
OF LOAD IMPEDANCE ON THE MASTER OSCILLATOR

The Master Oscillator and the first stage of the Master Amplifier were constructed as one unit. The Component values of the Amplifier were calculated by the use of the design equations in Appendix A and the results of these calculations are shown in Figure 3.21. The actual circuit of the Master Oscillator and the first stage of amplification is shown in Figure 3.21.



Q_{LD} = Q = 2N43A

R_e = 330 Ω

R_1 = 2M Ω

R_2 = 200,000 Ω

C_e = 4 μfd.

C_1 = 2 μfd.

T_f = 4 turns

T_c = 500 turns

T_1 = 500 turns

C_t = .006 μfd.

R = 33,000 Ω

R_{LD1} = 2500 Ω pot.

R_{LD2} = 22,000 Ω

R_{LD3} = 12,000 Ω

R_o = 120 Ω

C = .25 μfd.

T_{LD1} = 900 turns

T_{LD2} = 150 turns

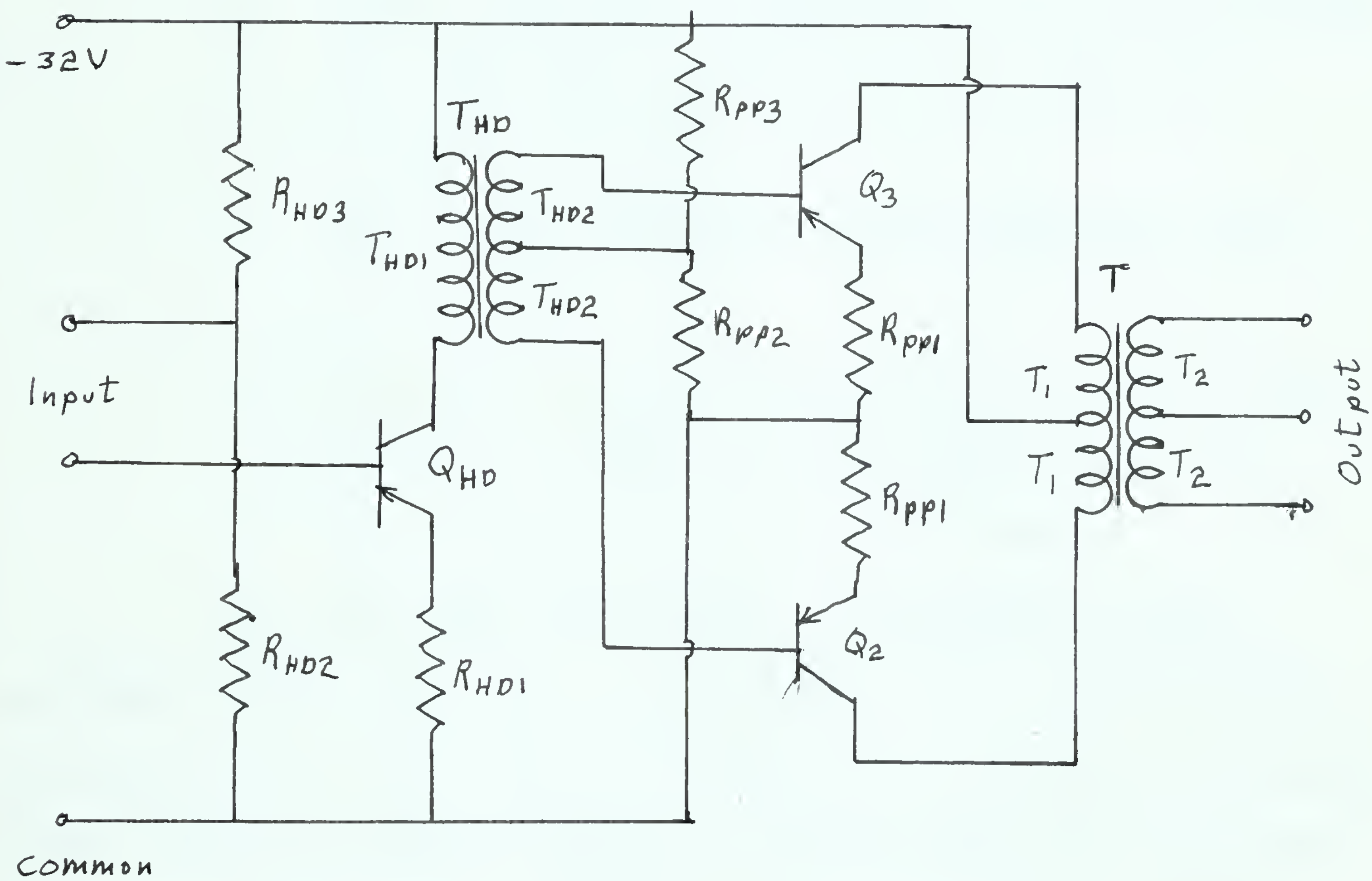
FIGURE 3.21

ACTUAL MASTER OSCILLATOR AND

EMITTER-FOLLOWER STAGE

3.10 MASTER AMPLIFIER

The design of the first stage of the Master Amplifier was included with the Master Oscillator. A gain adjustment was included in this stage. Figure 3.22 shows the remaining two stages of the Master Amplifier, which were built as one unit.



$R_{HD1} = 30\ \Omega$	$Q_{HD} = Q_2 = Q_3 = $ 2N540A
$R_{HD2} = 82\ \Omega$	$T_{HD1} = 100$ turns
$R_{HD3} = 795\ \Omega$	$T_{HD2} = 70$ turns
$R_{PP1} = 3\ \Omega$	For T_1 and T_2 see Table VII
$R_{PP2} = 10\ \Omega$	
$R_{PP3} = 300\ \Omega$	

FIGURE 3.22

MASTER AMPLIFIER

The component values shown in Figure 3.22 were obtained by the use of the equations in Appendix A and characteristic curves in Appendix B. Additional information concerning the design is listed below:

1. A stability factor of 3.5 was used.
2. Input impedance of the first stage was 60 ohms.
3. The impedance looking into the push-pull stage was 68 ohms.
4. The output impedance of the push-pull stage was 23 ohms.
5. The peak to peak voltage at the output of the push-pull stage was approximately 60 volts.
6. The maximum current was 1.2 amperes.

For the Power Transformer a Ferroxcube U-core (IF22B2-IF22B3) was used. According to the manufacturer's design charts, this U-core was adequate to handle the desired power and the required number of ampere turns for the transformer was 80. For a current of 1.2 amperes the possible number of turns was,

$$N = \frac{80}{1.2} = 66.7 \text{ turns}$$

Since the output peak to peak voltage was approximately 60 and a peak to peak voltage of approximately 4000 was needed, the turns ratio was calculated.

$$\text{Turns ratio} = \frac{4000}{66.7} = 60$$

The turns ratio of the primary of the Transformer was tapped at 66, 70 and 75 turns. The secondary was tapped at 4500 and 5300 turns. The possible turns ratio available are given in TABLE VII. It was found experimentally that the desired turns ratio was 60.

TABLE VII

POSSIBLE TURNS RATIOS OF POWER TRANSFORMER

T_2	T_1	T_2/T_1
4500	75	60.0
4500	70	64.3
4500	66	68.2
5300	75	70.7
5300	70	75.8
5300	66	80.3

The construction of this Transformer consisted of making the coils in two separate halves, in order to be able to obtain a center tap. The primary was wound with number 18 wire and the secondary was wound with number 36 wire.* A layer of tape was applied between each layer of wire.

* The Radio Amateur's Handbook, (American Radio Relay League, Inc., U.S.A.), 1960, p. 506

The heat sinks for the power transistors were made of pieces of copper plate 1/8 inch thick with a piece of 1/32 inch copper plate, in the form of a fin, on each side of the 1/8 inch copper plate. The size of the plate used was 3 by 9 inches. The heat sinks proved adequate.

3.11 RECTIFIER AND FINAL FILTER

Since the Power Transformer was hand wound it was convenient to include a center-tap on the output winding. As a result only two rectifiers were needed for full-wave rectification. The rectifiers had to withstand a maximum voltage of approximately 2000 volts. The rectifiers which were selected were the 1N1139, silicon type, of which the peak inverse voltage is specified as 3600 volts and the maximum D.C. current as 65 milliamperes.

For the smoothing filter it was decided to use a choke input in order to avoid large peak currents being drawn from the rectifier. The critical inductance, which is the minimum inductance to achieve continuous current through the rectifiers, is given as $\frac{R_L}{3\omega}$ *. Therefore, the critical inductance

*Millman Jacob, Vacuum-tube and Semiconductor Electronics, (McGraw-Hill Co., Inc., 1958) pp. 510-13.

for no load, when only the voltage divider is drawing current is,

$$L_{C_{N.L.}} = \frac{R_L}{3\omega} = \frac{915,000}{3 \times 2\pi \times 4000}$$

where, $915,000 = R_L$

4000 = frequency in c.p.s.

$$\underline{L_{C_{N.L.}} = 12.15 \text{ henries}}$$

At full-load the actual load impedance is 100,000 ohms. The effective load impedance including the voltage divider is,

$$R_{L_{F.L.}} = \frac{100,000 \times 915,000}{100,000 + 915,000} = 89,200\ \Omega$$

Therefore, the critical inductance at full load is,

$$L_{C_{F.L.}} = \frac{89.2}{3 \times 2\pi \times 4000}$$

$$\underline{L_{C_{F.L.}} = 1.18 \text{ henries}}$$

To complete the analysis of the smoothing filter, a Fourier analysis of a full-wave, rectified wave was done. Using the approximate component values of the actual filter, the amplitude of the ripple was estimated.

The Fourier series expansion expression is*,

$$f(\theta) = A_1 \sin\theta + A_2 \sin 2\theta + A_3 \sin 3\theta \ldots.$$

$$+ \frac{B_o}{2} + \underline{B_1} \cos\theta + B_2 \cos 2\theta + \ldots.$$

*Laws F.A., Electrical Measurements, (McGraw-Hill Co.,Inc. 1938), p.p. 677-78

The coefficients are given by,

$$A_n = \frac{1}{\pi}\int_0^{2\pi} f(\theta)\ \sin n\theta d\theta$$

$$B_n = \frac{1}{\pi}\int_0^{2\pi} f(\theta)\ \cos n\theta d\theta$$

The sine and cosine terms of the series can be combined. Thus,

$$A \sin \theta + B \cos \theta = \sqrt{A^2 + B^2}\ \sin\left(\theta + \tan^{-1}\frac{B}{A}\right)$$
$$= C \sin (\theta + \alpha')$$

Therefore, the original expression can be written as,

$$f(\theta) = \frac{B_o}{2} + C_1 \sin (\theta + \alpha'_1) + C_2 \sin (2\theta + \alpha'_2) + \ldots.$$

To aid the application of this expression to a full-wave rectified wave, Figure 3.23 was constructed.

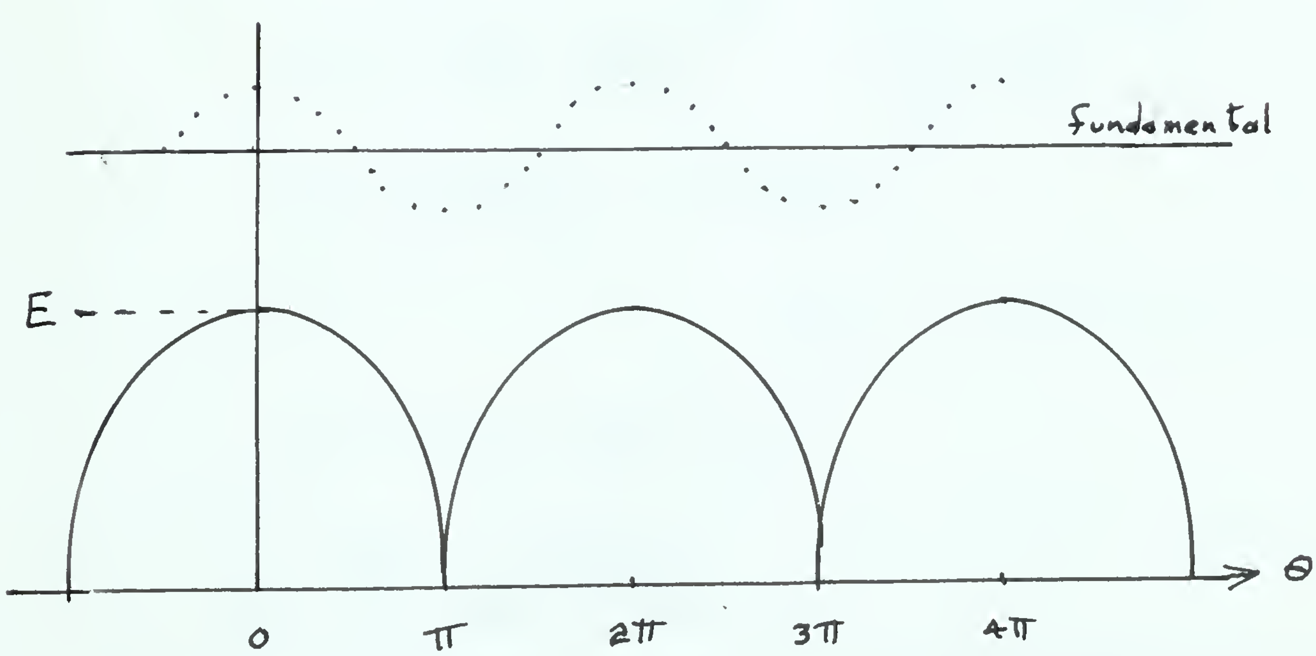


FIGURE 3.23

SKETCH OF A FULL-WAVE RECTIFIED WAVE

$$f(\theta) = E \cos \frac{\theta}{2} \qquad 0 < \theta < \pi$$

$$f(\theta) = -E \cos \frac{\theta}{2} \qquad \pi < \theta < 2\pi$$

Therefore,

$$B_o = \frac{E}{\pi} \int_0^{\pi} \cos \frac{\theta}{2}\, d\theta - \frac{E}{\pi} \int_{\pi}^{2\pi} \cos \frac{\theta}{2}\, d\theta$$

$$= \frac{2E}{\pi} \int_0^{\pi} \cos \frac{\theta}{2}\, d\theta = \frac{2E}{\pi} \left[2 \sin \frac{\theta}{2} \right]_0^{\pi}$$

$$\underline{B_o = \frac{4E}{\pi}}$$

$$B_n = \frac{E}{\pi} \int_0^{\pi} \cos \frac{\theta}{2} \cos n\theta d\theta - \frac{E}{\pi} \int_{\pi}^{2\pi} \cos \frac{\theta}{2} \cos n\theta\, d\theta$$

$$= \frac{2E}{\pi} \int_0^{\pi} \cos \frac{\theta}{2} \cos n\theta d\theta$$

Using identity, $\cos \alpha \cos B = 1/2 \left[\cos (\alpha - B) + \cos (\alpha + B) \right]$

$$B_n = \frac{E}{\pi} \int_0^{\pi} \left[\cos (n + 1/2)\theta + \cos (n - 1/2)\, \theta \right] d\theta$$

$$= \frac{E}{\pi} \left[\frac{\sin (n + 1/2)\theta}{n + 1/2} + \frac{\sin (n - 1/2)\theta}{n - 1/2} \right]_0^{\pi}$$

$$= \frac{E}{\pi} \left[\frac{(-1)^n}{n + 1/2} - \frac{(-1)^n}{n - 1/2} \right]_0^{\pi}$$

where $n = 1, 2, 3, \ldots.$

$$= \frac{E}{\pi} \quad (-1)^n \left[\frac{-1}{n^2 - 1/4} \right]$$

$$B_n = \frac{4E}{\pi} \quad \frac{(-1)^{n+1}}{(2n - 1)(2n + 1)}$$

Using this expression, the following coefficients were calculated:

$$B_1 = \frac{4}{3}\frac{E}{\pi}$$

$$B_2 = \frac{-4}{15}\frac{E}{\pi}$$

$$B_3 = \frac{4}{35}\frac{E}{\pi}$$

$$B_4 = \frac{-4}{63}\frac{E}{\pi}$$

The calculations for A_n are,

$$A_n = \frac{E}{\pi}\int_0^{\pi} \cos\frac{\theta}{2}\sin n\theta d\theta - \frac{E}{\pi}\int_0^{2\pi} \cos\frac{\theta}{2}\sin n\theta d\theta$$

where, $1/2 \sin\alpha \cos B = \sin(\alpha + B) + \sin(\alpha - B)$

$$A_n = \frac{E}{2\pi}\int_0^{\pi}\left[\sin(n + 1/2)\,\theta + \sin(n - 1/2)\,\theta\right] d\theta$$

$$- \frac{E}{2\pi}\int_{\pi}^{2\pi}\left[\sin(n + 1/2)\theta + \sin(n - 1/2)\,\theta\right] d\theta$$

$$= \frac{-E}{2\pi}\left\{\left[\frac{\cos\,(n + 1/2)\theta}{n + 1/2} + \frac{\cos(n - 1/2)\theta}{n - 1/2}\right]_0^{\pi} - \left[\frac{\cos(n + 1/2)\theta}{n + 1/2} + \frac{\cos(n - 1/2)\theta}{n - 1/2}\right]_{\pi}^{2\pi}\right\}$$

$$= \frac{-E}{2\pi}\left[\frac{-1}{n + 1/2} - \frac{1}{n - 1/2} - \frac{-1}{n + 1/2} - \frac{1}{n - 1/2}\right]$$

$$A_n = 0$$

Therefore, the general expression can be written as,

$$f(\theta) = \frac{B_o}{2} + \sqrt{A_1^{\,2} + B_1^{\,2}}\sin\left(\theta + \tan^{-1}\frac{B_1}{A_1}\right) + \sqrt{A_2^{\,2} + B_2^{\,2}}$$

$$\sin\left(2\theta + \tan\frac{B_2}{A_2}\right) + \ldots.$$

Since A_n is zero the terms $\frac{B_n}{A_n}$ are infinite. Therefore,

$$\tan^{-1} \frac{B_n}{A_n} = \pi/2$$

Substituting this into the general expression and inserting the proper coefficients, the following result was obtained:

$$f(\theta) = \frac{2E}{\pi} + \frac{4}{3}\frac{E}{\pi} \sin\left(\theta+\frac{\pi}{2}\right) - \frac{4}{15}\frac{E}{\pi} \sin\left(2\theta+\frac{\pi}{2}\right) +$$

$$\frac{4}{35}\frac{E}{\pi} \sin\left(3\theta+\frac{\pi}{2}\right) - \frac{4}{63}\frac{E}{\pi} \sin\left(4\theta+\frac{\pi}{2}\right) + \ldots.$$

where $\theta = \omega t$ of the fundamental.

Rewriting the above expression, the result is,

$$f(\theta) = \frac{2E}{\pi} + \frac{4}{3}\frac{E}{\pi} \cos \omega t - \frac{4}{15}\frac{E}{\pi} \sin 2(\omega t + 45) +$$

$$\frac{4}{35}\frac{E}{\pi} \sin 3(\omega t + 30) - \frac{4}{63}\frac{E}{\pi} \sin 4(\theta + 22.5)$$

The D.C. component of this wave is,

$\underline{E_{DC} = \frac{2E}{\pi}}$ where E is the maximum voltage.

Writing the harmonics in terms of the D.C. component the following results were obtained:

1. Fundamental $= \frac{2}{3} E_{DC} \cos \omega t$
2. Second harmonic $= -\frac{2}{15} E_{DC} \sin 2(\omega t + 45)$
3. Third harmonic $= \frac{2}{35} E_{DC} \sin 3(\omega t + 30)$
4. Fourth harmonic $= -\frac{2}{63} E_{DC} \sin 4(\omega t + 22.5)$

After having the wave Fourier analyzed the next step was to see what happened to each harmonic as it passed through the filter.

Figure 3.24 illustrates the filter.

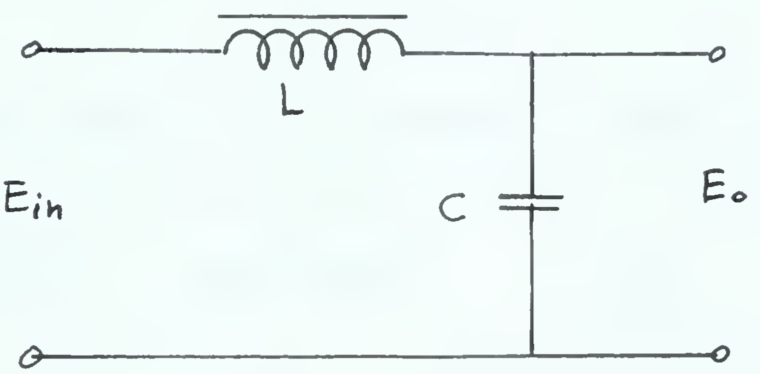


FIGURE 3.24

CHOKE INPUT FILTER

Since the reactance of capacitor was small compared to the load resistance, the A.C. current through the capacitor was,

$$\frac{E_{in}}{\omega L - \frac{1}{\omega C}}$$

It followed that the A.C. voltage across the capacitor and load was

$$\frac{E_{in}}{(\omega L - \frac{1}{\omega C})\omega C} = \frac{E_{in}}{\omega^2 LC - 1}$$

Therefore, the ratio of the A.C. output voltage to the A.C. input voltage was,

$$\frac{1}{\omega^2 LC - 1} \quad *$$

TABLE VIII shows the resulting A.C. voltages across the output of the filter in terms of the D.C. component, using an inductance of 25 henries and a capacity of 10 μfd.

TABLE VIII

A.C. VOLTAGES SEEN AT THE OUTPUT OF REGULATED SUPPLY

Harmonic	Freq. Rod/Sec.	Amplitude	$\frac{AC\ E_o}{AC\ E_{in}}$	Peak Voltage of Harmonic
first	8,000	2/3 E_{DC}	1.587×10^{-6}	$1.059 \times 10^{-6} E_{DC}$
second	16,000	2/15 E_{DC}	$.397 \times 10^{-6}$	$.053 \times 10^{-6} E_{DC}$
third	24,000	2/35 E_{DC}	$.1763 \times 10^{-6}$	$.010 \times 10^{-6} E_{DC}$
fourth	32,000	2/63 E_{DC}	$.0992 \times 10^{-6}$	$.003 \times 10^{-6} E_{DC}$

By adding the amplitudes of the four harmonics directly the sum is $1.125 \times 10^{-6} E_{DC}$. In other words, the peak to peak voltage of the maximum possible ripple is 0.000225 per cent of the D.C. output. This is a favourable result.

The capacitor obtained for the filter was a 10 μfd., 2000 volt, oil-filled type. The inductance was obtained from

*Terman F.E., Electronic and Radio Engineering, (McGraw-Hill Co., Ltd., N.Y., 1955) p.p. 721-22.

a hand-wound coil on two pairs of Ferroxcube cores (1F22B2 - 1F22B3).

In winding the choke, the object was to obtain an inductance as high as possible. The B-H curve for the material used was given by the manufacturer and is shown in Figure 3.25.

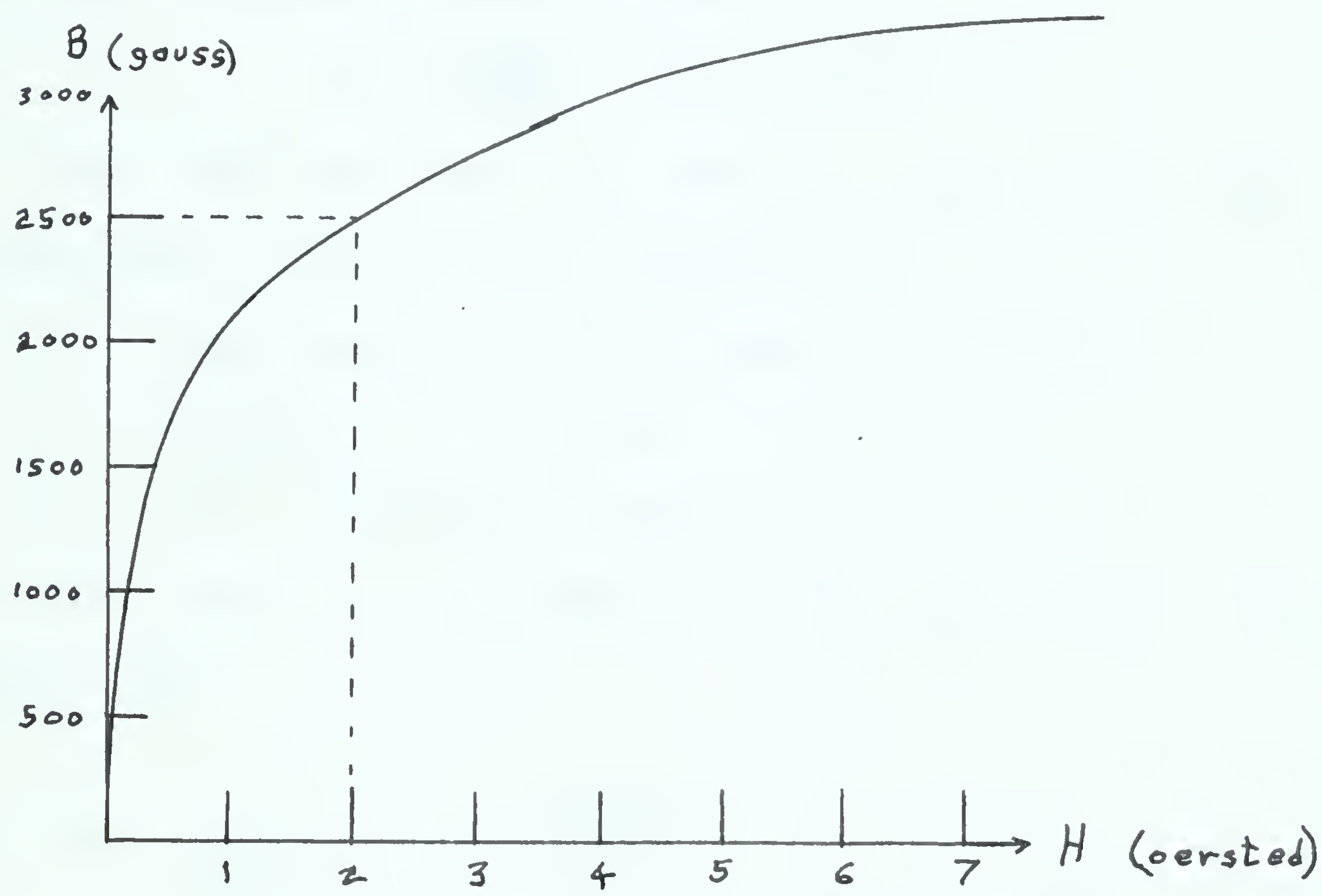


FIGURE 3.25

B-H CURVE FOR THE CHOKE CORE MATERIAL
(1F22B2 - 1F22B3)

After inspecting the curve, it was decided to operate at approximately 2500 gauss and 2 oersteds. Converting two oersteds to ampere-turns per centimeter, the result was,

$$2 \text{ oer.} = \frac{2 \times 10^3}{4\pi \, 10^2} \text{ NI/cm.}$$

$$= 1.6 \text{ NI/cm.}$$

The mean length of the core was determined as 19.3 centimeters. Therefore, the possible ampere-turns was,

$$19.3 \times 1.6 = 31 \text{ NI}$$

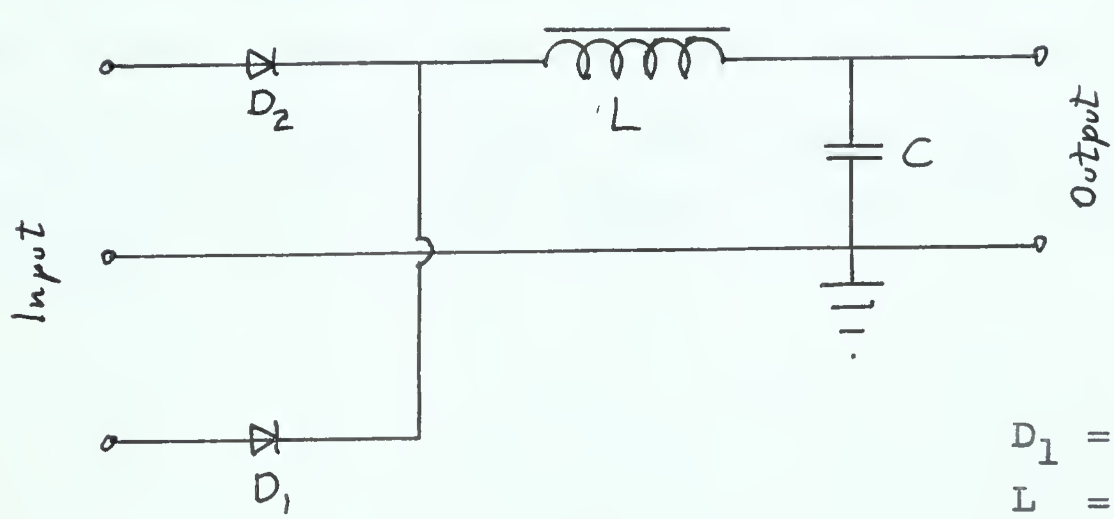
Since the maximum D.C. current was approximately 11 milliamperes, the allowable number of turns was,

$$N = \frac{31}{0.011} = 2820 \text{ turns}$$

The choke was wound and 2700 turns were used. The resulting choke constants are listed below:

1. Inductance = 25.8 hys. (at 1000 cps.)
2. Q-factor = 135
3. D.C. resistance = 200 Ω.

The circuit of the actual Rectifier and Filter is shown in Figure 3.26.



$D_1 = D_2 = 1N1139$
$L = 25.8$ hy.
$C = 10$ μfd.

FIGURE 3.26

ACTUAL RECTIFIER AND FILTER CIRCUIT

3.12 D.C. SUPPLIES

Three stable D.C. supplies were needed to supply the necessary D.C. power. As far as the power required, two supplies would have been sufficient, but due to grounding problems three supplies were needed. The design for these supplies was obtained from a supply already built. As a result, only the circuit diagrams are shown. Figure 3.27 shows the circuit of the supply used.

The power supply which supplied the Master Amplifier needed a current output of approximately 1.1 amperes. Since this was more than the supply in Figure 3.27 could produce, a 210M transformer had to be used instead of the 167Q transformer. The rest of the circuit remained unaltered.

In the power supply which was to drive the zener diodes, the voltage dividing resistors used were wire-wound instead of the carbon type. This was done to achieve greater temperature stability. Also, the circuit was altered slightly. The circuit diagram of this supply is shown in Figure 3.28.

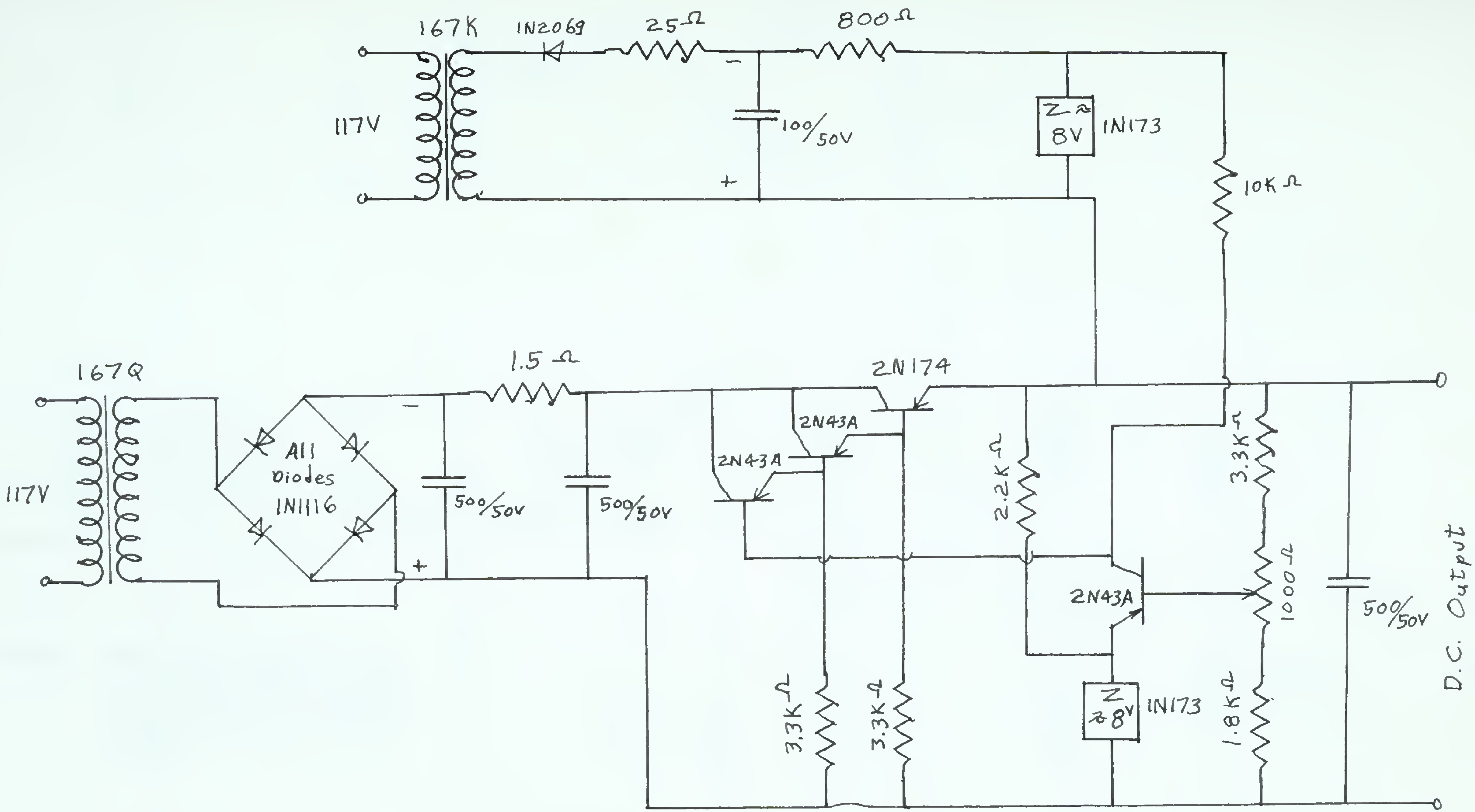
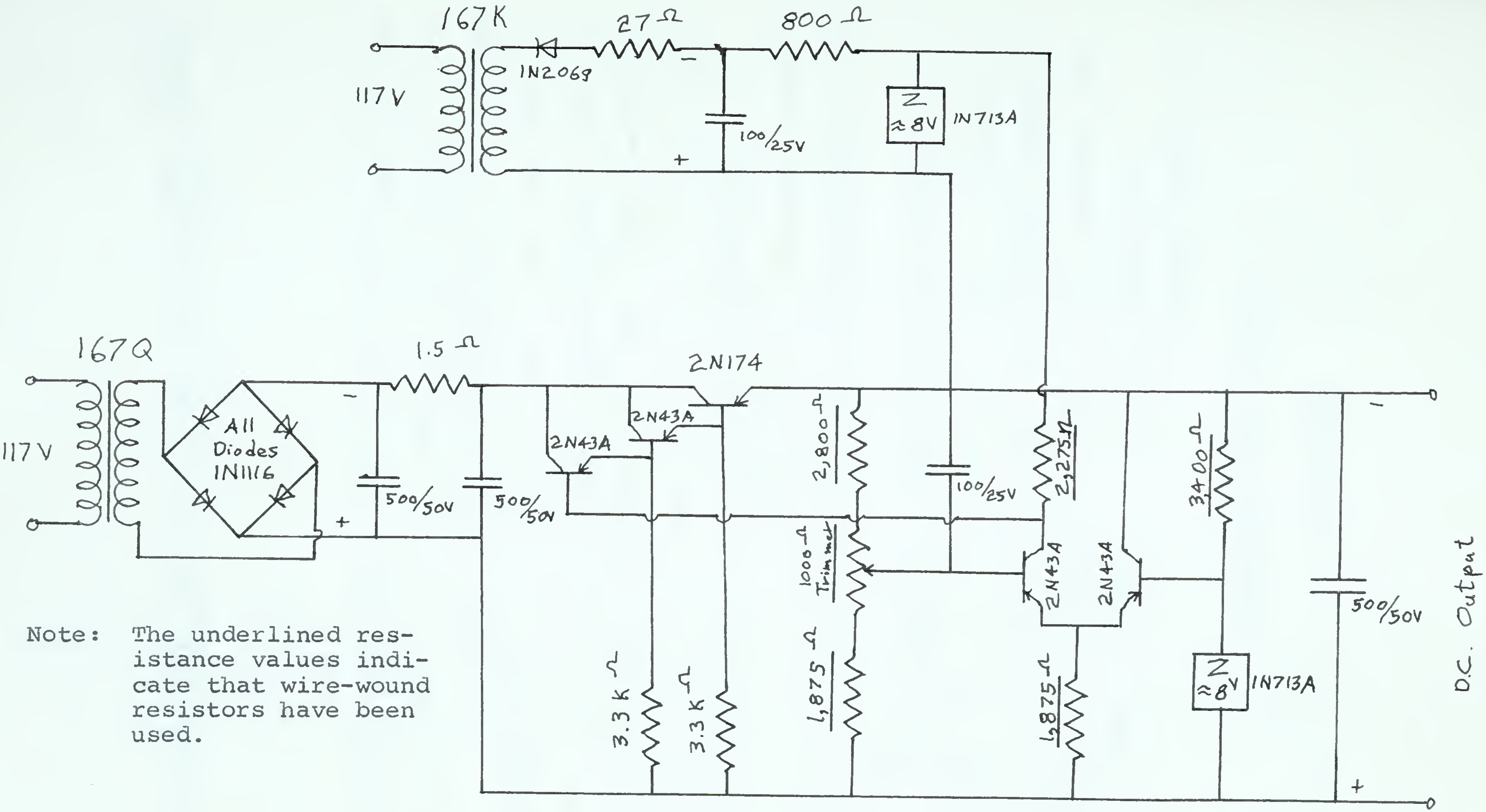


FIGURE 3.27

STABLE D.C. SUPPLY (0.8 A. and 25 V.)



Note: The underlined resistance values indicate that wire-wound resistors have been used.

FIGURE 3.28

STABLE D.C. SUPPLY FOR REFERENCE VOLTAGE (0.8 A. and 25 V.)

CHAPTER IV

ANALYSIS AND CONCLUSION

4.1 SYSTEM STABILITY

To determine whether the system was stable the open-loop frequency response was obtained and the log-magnitude and phase diagrams made. From these the system stability was observed. The steady state error of a type zero system with a constant actuating signal is given by $\frac{R_o}{1 + K_o}$ * where R_o is the reference or actuating signal and K_o is the loop gain or position error coefficient.

The frequency response of the inner loop and of the major loop was obtained at no load and at full load of the Regulated Supply. The data obtained for the inner loop and no load is shown in TABLE IX and the data for the inner loop at near full load is shown in TABLE X. The log-magnitude and phase plots of this data are shown in Figure 4.1.

The data for the major loop at no load is in TABLE XI and the data for the major loop, at near full load is in

*D'Azzo J. J., Houpis C. H., Feedback Control System Analysis and Synthesis, (McGraw-Hill Co., Inc., N.Y., 1960), p. 116.

TABLE XII. The major loop frequency response was obtained with the inner loop in operation. The log-magnitude and phase plots of this data are shown in Figure 4.2 and Figure 4.3, respectively.

Inspecting the open-loop frequency response plots, it is seen that the log-magnitude curves, in all cases, cross the zero decibel line at a slope greater than -40 decibels per decade. Also, when the log-magnitude curve crosses the zero decibel line the phase angles, in all cases, are more than -180 degrees. These conditions indicate that the system is stable. It could be noted that since the observed curves did not follow the -20 or -40 decibels per decade slopes,the system must have possessed nonlinearities.

With only the inner loop closed, it was found that the output voltage varied approximately 2 per cent when the load current varied from zero to full load. If 2 per cent of the output voltage was considered as the reference which was actuated by a step load current, then R_o can be made equal to 0.02 V_o where V_o is the output voltage. From the log-magnitude plot in Figure 4.3 the value of K_o is 50 decibels or 318, numerically. The value of K_o was obtained from the full load plot since it had the lower gain and was more critical. Therefore, the steady state error is,

$$= \frac{0.02\ V_o}{1 + 318} = 0.0063\ \%\ \text{of}\ V_o.$$

This value is within the limits of the requirement.

TABLE IX

OPEN LOOP FREQUENCY RESPONSE OF INNER LOOP AT NO LOAD

(V_{out} = 900 V and I_{out} = 0.0 ma)

Frequency		V_{out}	V_{in}	V_{out}/V_{in}		Phase
c.p.s.	rad/sec.	V_{pp}	V_{pp}	Ratio	db	deg.
.2	1.256	1.6	2.80	.572	- 4.8	+ 12
.5	3.14	1.7	2.85	.596	- 4.5	+ 5
1	6.28	1.7	2.85	.596	- 4.5	+ 1
2	12.56	1.7	2.85	.596	- 4.5	0
4	25.1	1.7	2.85	.596	- 4.5	- 3
7	44.0	1.65	2.85	.579	- 4.75	- 4
10	62.8	1.6	2.85	.562	- 5.0	- 5
20	125.6	1.55	2.85	.544	- 5.3	- 6.5
40	251	1.5	2.80	.536	- 5.4	- 11
70	440	1.45	2.75	.528	- 5.6	- 19
100	628	1.4	2.70	.518	- 5.7	- 29
200	1256	1.16	2.30	.504	- 6.0	- 51
400	2510	.8	1.65	.484	- 6.3	- 82
700	4400	.4	1.08	.370	- 8.7	-114
1000	6280	.2	.64	.313	-10.2	-127

TABLE X

OPEN LOOP FREQUENCY RESPONSE OF INNER LOOP

AT NEAR FULL LOAD

(V_{out} = 900 V and I_{out} = 9.5 ma)

Frequency		V_{out}	V_{in}	V_{out}/V_{in}		Phase
c.p.s.	rod/sec.	V_{pp}	V_{pp}	Ratio	db	deg.
.2	1.256	2.5	2.9	.862	+1.8	+ 15
.5	3.14	3.2	2.9	1.103	+1.0	+ 13
1	6.28	3.4	2.9	1.172	+1.3	+ 5
2	12.56	3.4	2.9	1.172	+1.3	0
4	25.1	3.2	2.9	1.103	+1.0	- 5
7	44.0	3.1	2.9	1.07	+ .6	- 7
10	62.8	2.8	2.9	.966	- .3	- 10
20	125.6	2.5	2.85	.877	-1.2	- 12
40	251	2.4	2.85	.843	-1.5	- 15
70	440	2.3	2.8	.822	-1.7	- 23
100	628	2.2	2.7	.815	-1.8	- 30
200	1256	1.9	2.35	.808	-1.9	- 53
400	2510	1.3	1.68	.774	-2.3	- 90
700	4400	.8	1.08	.740	-2.7	-128
1000	6280	.5	.78	.641	-3.9	-156

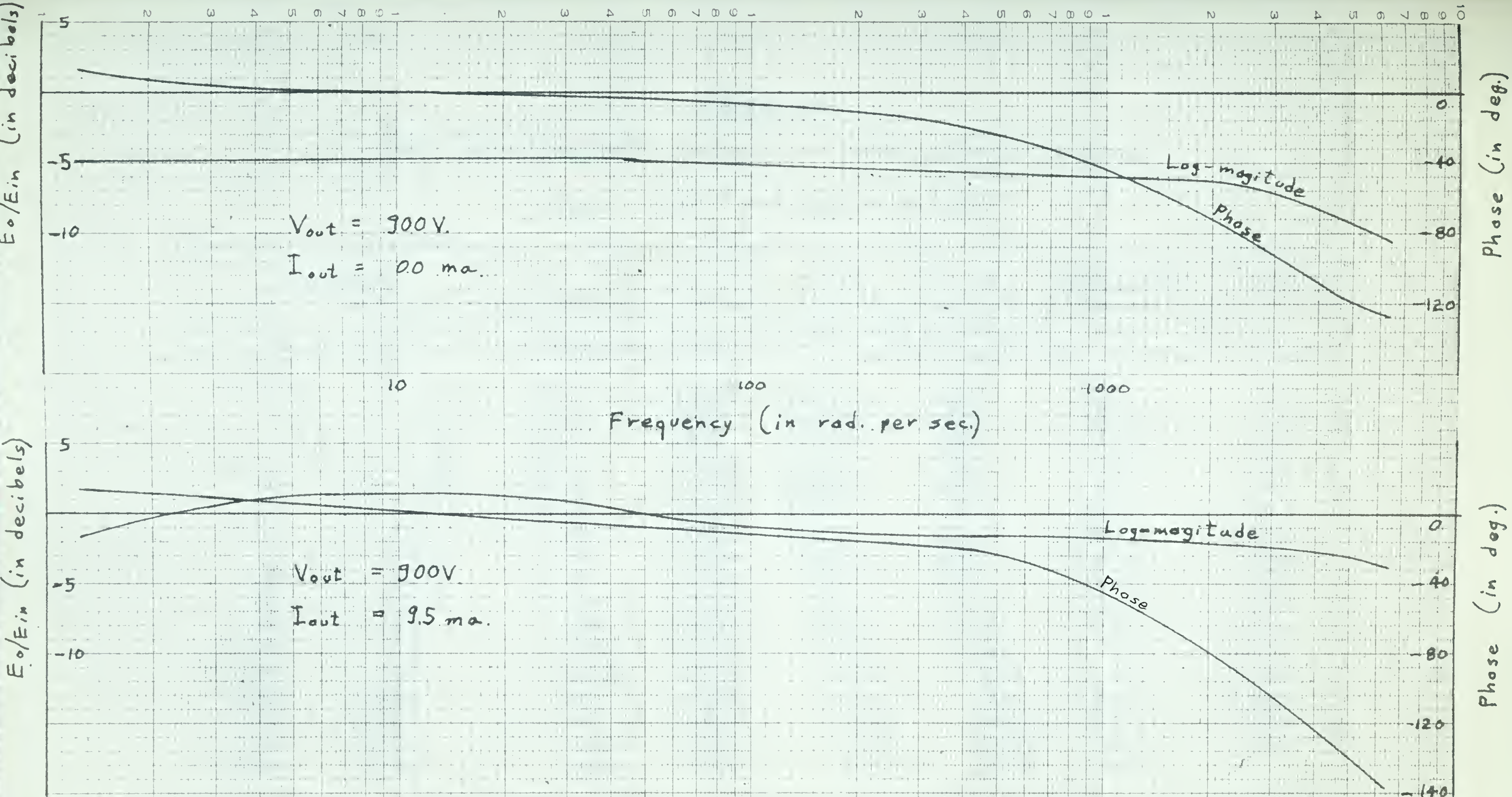


FIGURE 4.1

OPEN LOOP FREQUENCY RESPONSE OF INNER LOOP

TABLE XI

OPEN LOOP FREQUENCY RESPONSE OF MAJOR LOOP AT NO LOAD

(V_{out} = 950 V and I_{out} = 0.0 ma)

Frequency		V_{out}	V_{in}	V_o/V_{in}	V_o/V_{in} (Adjusted)		Phase
c.p.s.	rad/sec.	V_{p-p}	V_{p-p}	(Measured)	Ratio	db	deg.
.1	.628	40.0	.0008	50,000	850	58.6	- 33
.2	1.256	100.0	.002	50,000	850	58.6	- 33
.4	2.51	21.0	.001	21,000	357	51.0	- 72
.7	4.40	26.5	.0016	16,500	282	49.0	- 83
1	6.28	32.0	.0032	10,000	170	44.5	- 87
2	12.56	16.6	.0034	4,880	83	38.4	- 94
4	25.1	7.4	.0034	2,175	36.9	31.3	-101
7	44.0	3.8	.0034	1,120	19	25.5	-103
10	62.8	2.4	.0034	706	12	21.5	-104
20	125.6	1.34	.0034	394	6.7	16.5	-107
40	251	.54	.0034	159	2.7	8.5	-114
70	440	.27	.0036	75	1.28	2.2	-125
100	628	.165	.0036	45.8	.778	- 2.2	-135
200	1256	.06	.0036	16.7	.284	- 11.0	-157
400	2510	.02	.0038	5.26	.0894	- 21.0	-194
700	4400	.008	.0038	2.1	.0367	- 28.7	-200
*1000	6280	.004	.0038	1.05	.0178	- 35.0	

* The phase angle could not be obtained due to 60 cycle interference.

Note: V_o/V_{in} adjustment is due to Voltage Divider, since the measured V_o was across the output and the amount fed was a portion of this.

TABLE XII

OPEN LOOP FREQUENCY RESPONSE OF MAJOR LOOP AT NEAR FULL LOAD

(V_{out} = 950 V and I_{out} = 8.0 ma)

Frequency		V_{out}	V_{in}	V_o/V_{in}	V_o/V_{in} (Adjusted)		Phase
c.p.s.	rad/sec.	(p-p)	(p-p)	(measured)	Ratio	db	deg.
.1	.628	11.2	.0006	18,700	318	50.0	- 2.0
.2	1.256	36.0	.0018	20,000	340	51.6	- 5.5
.5	3.14	45.0	.002	22,500	382	51.7	- 19
1	6.28	42.0	.0018	23,300	396	51.8	- 45
2	12.56	30.0	.0018	16,700	283	49.0	- 85
4	25.1	11.0	.002	5,500	93.5	39.4	-110
7	44.0	4.7	.002	2,350	40.0	32.0	-114
10	62.8	2.9	.0022	1,320	22.4	27.0	-115
20	125.6	1.26	.0022	573	9.73	19.8	-115
40	251	.58	.0022	264	4.49	13.0	-120
70	440	.28	.0022	127.3	2.165	6.7	-133
100	628	.16	.0022	72.7	1.236	1.8	-143
200	1256	.055	.0022	25.0	.425	- 7.5	-164
400	2510	.02	.0022	9.1	.155	-16.2	-190
700	4400	.007	.0022	3.18	.0541	-25.3	-120
* 1000	6280	.003	.0022	1.364	.0232	-32.7	

* The phase angle could not be obtained because of 60 cycle interference.

Note: V_o/V_{in} adjustment is due to Voltage Divider, since the measured V_o was across the output and the amount fed was a portion of this.

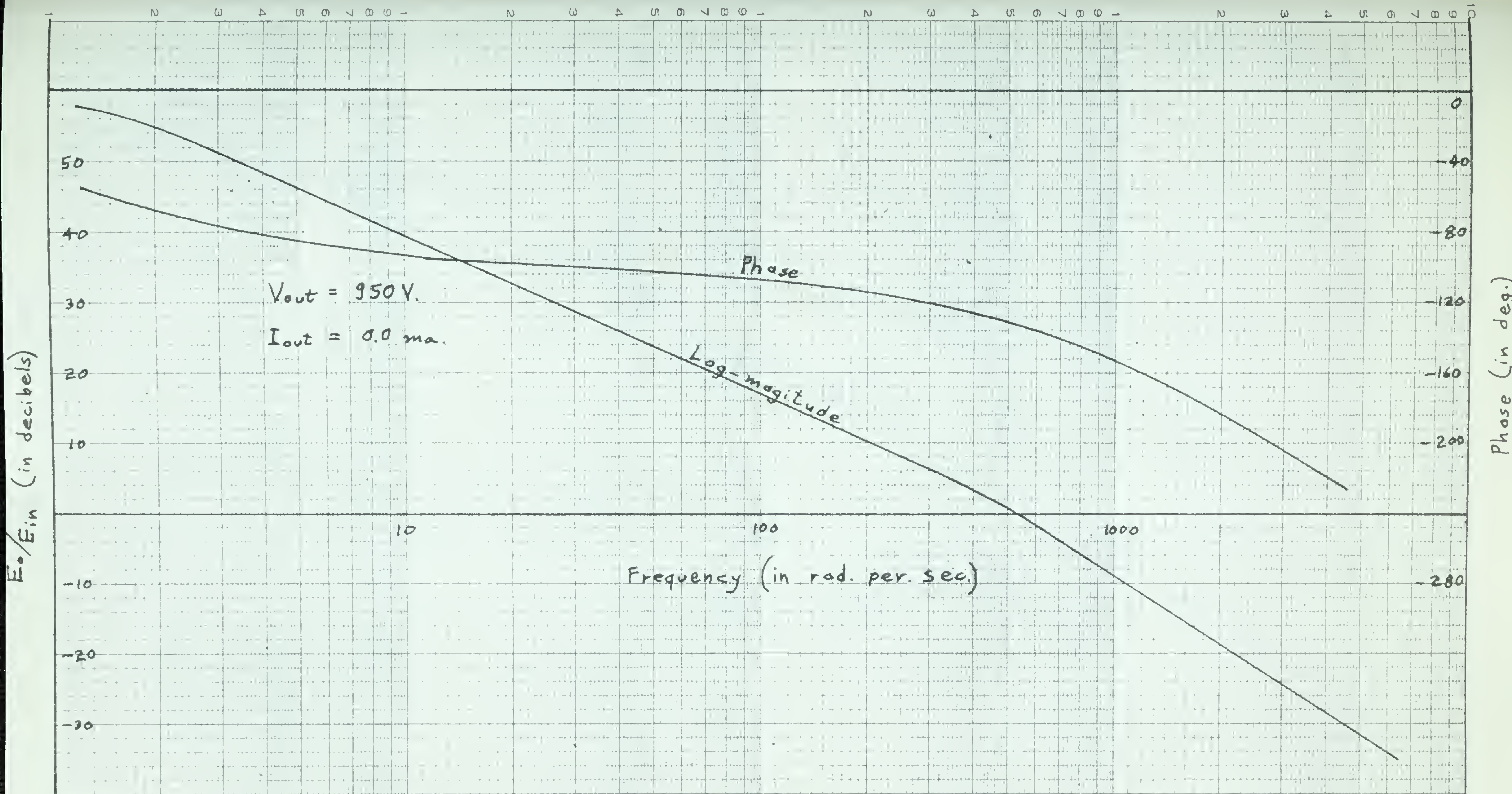


FIGURE 4.2

OPEN LOOP FREQUENCY RESPONSE OF MAJOR LOOP

AT NO LOAD

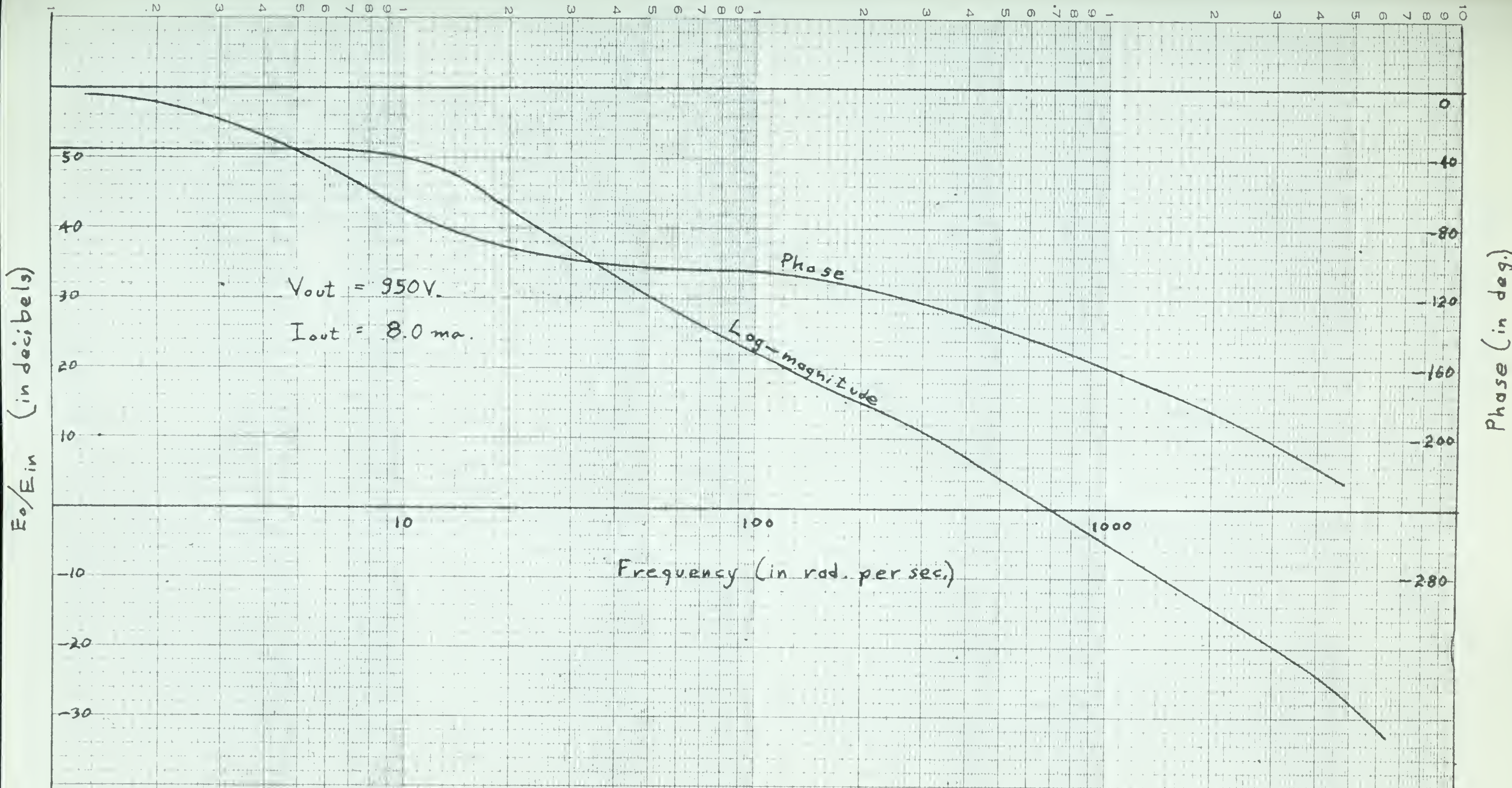


FIGURE 4.3

OPEN LOOP FREQUENCY RESPONSE OF MAJOR LOOP

AT NEAR FULL LOAD

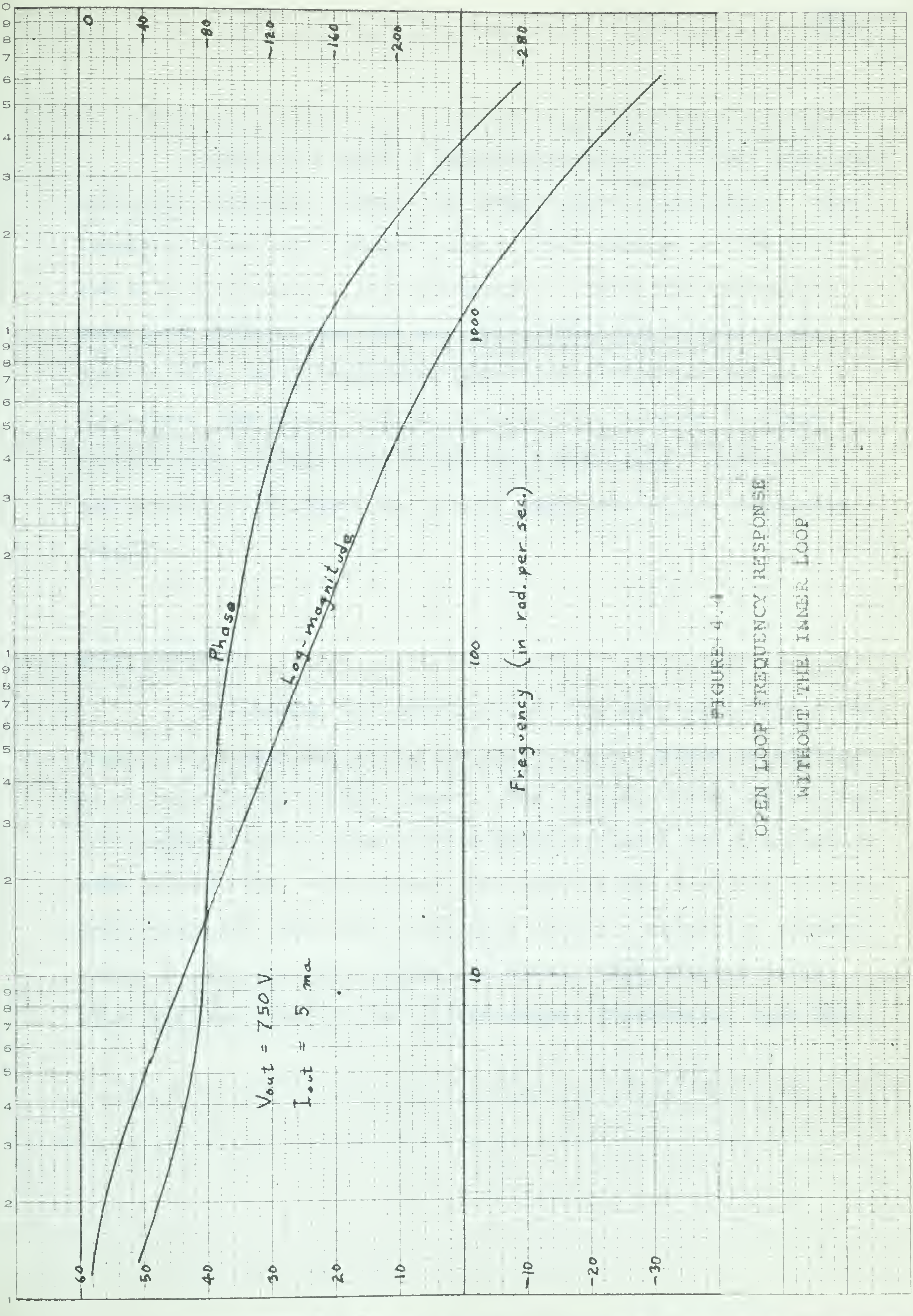


FIGURE 4.4

OPEN LOOP FREQUENCY RESPONSE

WITHOUT THE INNER LOOP

Figure 4.4 shows a log-magnitude plot of the open-loop frequency response without the inner loop in operation. This response curve was obtained with a load voltage of 750 volts and a load current of 5 milliamperes. Since the log-magnitude plot follows the -20 and -40 decibel per decade slope rather well, it is concluded that this system was linear. Therefore, the inner loop which provided positive feedback proportional to the load caused nonlinearities. However, it was needed to decrease the output impedance of the Regulated Supply.

4.2 RESULTS

To measure the regulation of the Regulated Supply the output was inspected at the Voltage Divider, since it appeared at a lower level at this point. The 17 volts from the Divider was compared to 17 volts from a battery source and the difference between the voltages was measured as the load was varied. The reason for this was to be able to use a sensitive voltage scale to measure the voltage variation. The voltage variation was measured with an oscilloscope, (Tektronix, type 502),

in the most sensitive range, which was 2 millivolts per division. The scope was D.C. coupled. The results are shown in TABLE XIII.

TABLE XIII

STATIC VARIATION OF OUTPUT VOLTAGE

OF REGULATED SUPPLY

Load	Voltage Variation
No load	0.0 mv.
Half load	0.5 mv.
Full load	0.6 mv.

The maximum voltage variation due to load, of the Regulated Supply, was 0.6 millivolts in 17 volts or 0.0035 per cent. This corresponds to 35 millivolts variation in 1000 volts.

Observing the output of the Regulated Supply on an oscilloscope, (Tektronix, type 502), it was seen that the 8000 cycle per second ripple was 5 millivolts peak to peak. Also, there was a 60 cycle component of 6 millivolts, peak to peak, which probably was caused by stray pick-up.

Figure 4.5 shows the response of the Regulated Power Supply to a step load. The response was obtained for a step, from no load to full load, and from full load to no load. This was done by switching the load current and potographing the output voltage variation on an oscilloscope, (Textronix, type 502).

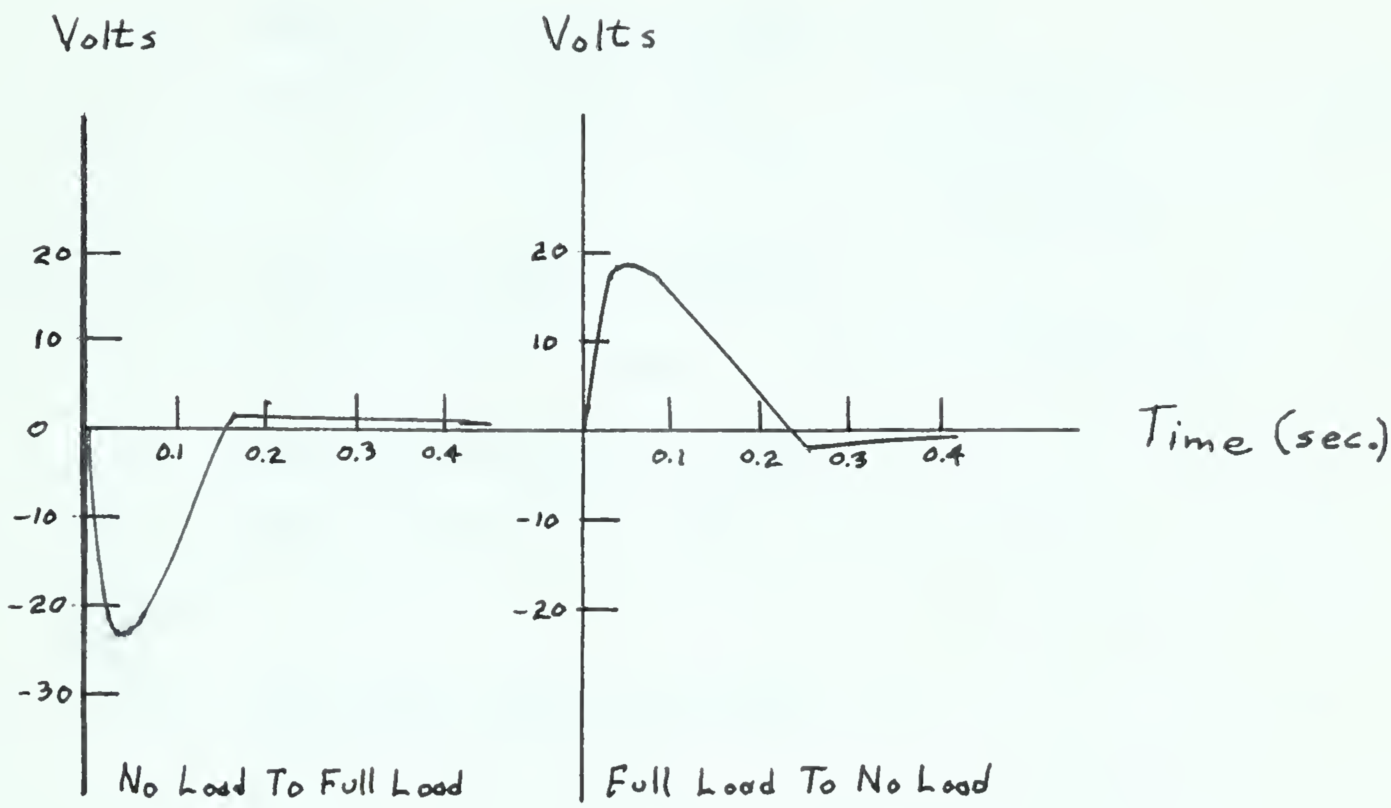


FIGURE 4.5

STEP-RESPONSE OF REGULATED SUPPLY

The list below summarizes the results obtained from the Regulated Supply:

1. Maximum steady state error = 35 mv.
2. 8000 c.p.s. ripple = 0.5 $mv._{p-p}$
3. 60 c.p.s. ripple = 0.6 $mv._{p-p}$
4. Total voltage variation from no load to full load = 36.1 mv. = 0.0036 %.

5. Maximum transient from
   no load to full load = 24 V.
   = 2.4 %.
6. Settling time, from no
   load to full load = 0.16 sec.
7. Maximum transient from
   full load to no load = 18 V.
   = 1.8 %.
8. Settling time, from no
   load to full load = .24 sec.

From the above list it can be seen that the voltage variation from no load to full load is 0.0036 per cent and the required regulation was 0.01 per cent. Therefore, the system is well within the specified limits of regulation.

4.3 CONCLUSION

It is considered that this work has shown that high voltage, well regulated, D.C. supplies can be built with the use of semiconductors.

Even though the Regulated Supply operated well some improvements are possible. The output impedance of the final common-emitter stage of the Master Amplifier was 23 ohms and with a transformer ratio of 60 the output impedance, with no current feedback was,

$$= (60)^2 \times 23 = 82,800\ \Omega.$$

This value of output impedance could have been reduced considerably if an emitter-follower configuration, for the final stage of the Master Amplifier, was used. Also, the range of the Demodulator could have been increased, since its output was limited to a maximum voltage swing of approximately 3 volts. With these two improvements and possibly an increased gain of the Error Amplifier, the current feedback loop could have been eliminated and this would have simplified the overall system.

The Master Oscillator design could have been altered so that the by-pass capacitors used could have been made smaller or eliminated entirely. This would have improved the frequency response of the system.

Additional problems could have been investigated with the aid of more specialized equipment. Operation in different ambient temperatures could have been a problem of interest if ovens were available. Also, long time variation, or drift, could have been investigated but in this case voltage-measuring devices of approximately 0.001 per cent accuracy or better would have been necessary. As a result, the problem was limited mainly to regulation.

The development of the semiconductorized, Regulated Supply is considered as successful. The knowledge gained was intense, both from the theoretical and practical point of view.

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APPENDIX A

AMPLIFIER BIASING CIRCUIT

The configuration adopted for amplifier biasing is shown in Figure A.1. The equivalent circuit of Figure A.1 is shown in Figure A.2.*

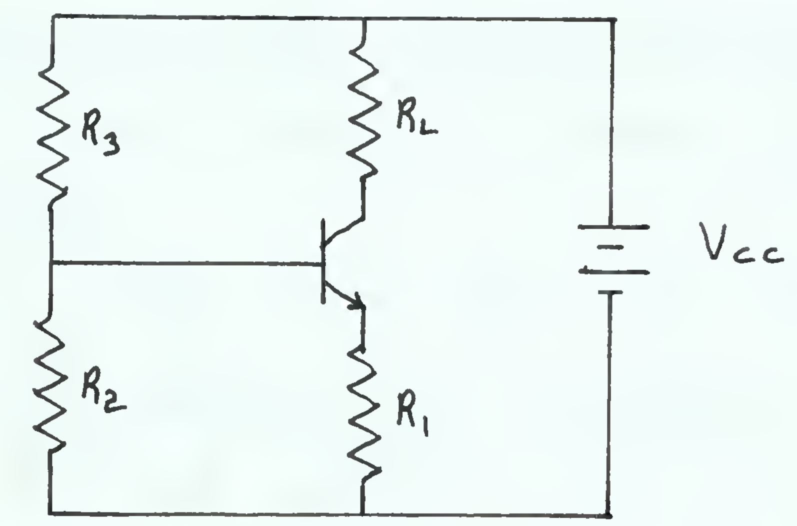

FIGURE A.1

BIASING ARRANGEMENT

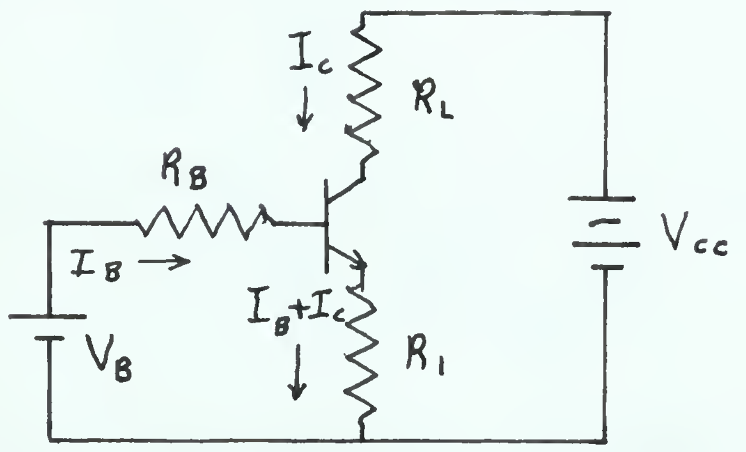

FIGURE A.2

EQUIVALENT CIRCUIT OF BIASING ARRANGEMENT

*Millman J., Vacuum-tube and Semiconductors Electronics, (McGraw-Hill Co., Ltd., N.Y., 1958), p.234

The equivalent circuit shown in Figure A.2 was obtained by replacing Figure A.1 by its Thevenin equivalent. Where,

$$V_B = \frac{R_2\ V_{CC}}{R_2 + R_3} \qquad \text{A.}$$

$$R_B = \frac{R_2 R_3}{R_2 + R_3} \qquad \text{B.}$$

Resistance R_B is the effective resistance looking back from the base terminal. Kirchoff's voltage law around the base circuit yields,

$$-V_B + I_B R_B + (I_B + I_C)R_1 + V_{BE} = 0 \qquad \text{C.}$$

Since I_B is normally small as compared to I_C, it can be neglected. Therefore, equation C can be written as,

$$I_B = \frac{V_B}{R_B} - \frac{I_C\ R_1 + V_{BE}}{R_B} \qquad \text{D.}$$

From equations A and B,

$$\frac{V_B}{R_B} = \frac{V_{CC}}{R_C} \ .$$

Using this equation and equation D it follows that,

$$I_B = \frac{V_{CC}}{R_3} - \frac{I_C R_1 + V_{BE}}{R_B} \qquad \text{E.}$$

Kirchoff's law around the collector circuit yields,

$$-V_{CC} + I_C(R_L + R_1) + I_B\ R_1 + V_{CE} = 0.$$

Therefore,

$$R_1 = \frac{(V_{CC} - V_{CE} - I_C R_L)}{I_C + I_B} \qquad \text{F.}$$

The stability factor, which is defined as dI_c/dI_{co}, for this configuration is,

$$S = \frac{1}{1 - \alpha + \alpha \left(\frac{R_1}{R_1 + R_B}\right)} \quad * \qquad G.$$

* Millman J., Vacuum-tube and Semiconductor Electronics, (McGraw-Hill Co., Inc., N.Y., 1958)

APPENDIX B

HYBRID PARAMETERS AND CHARACTERISTICS OF TRANSISTORS

The hybrid parameters and characteristic curves were needed for the transistors 2N43A, 2N540A,and 2N118. The parameters were obtained for the common-emitter configuration, by the use of the Transistor Curve-Tracer. The parameters for the 2N43A transistor are listed below,(for I_c = 1 ma., and V_c = -5 V.), and the common-emitter characteristics are shown in Figure B.1.

1. h_{11} = 1246 ohms
2. h_{12} = 22.4×10^{-5}
3. h_{21} = 43.5
4. h_{22} = 22.3×10^{-6} mhos

The parameters for the 2N540A transistor are listed (for I_c = 0.6 amps. and V_c = -35 V.). The common-emitter characteristics are shown in Figure B.2.

1. h_{11} = 69 ohms
2. h_{12} = 10×10^{-4}
3. h_{21} = 92.4
4. h_{22} = 4.0×10^{-3} mhos

The parameters for the transistor 2N118 are listed below,(for I_c = 5 ma. and V_c = -7 V.), and the common-emitter

characteristics are shown in Figure B.3.

1. $h_{11} = 4500$ ohms
2. $h_{12} = 5 \times 10^{-4}$
3. $h_{21} = 36.7$
4. $h_{22} = 20 \times 10^{-6}$ mhos.

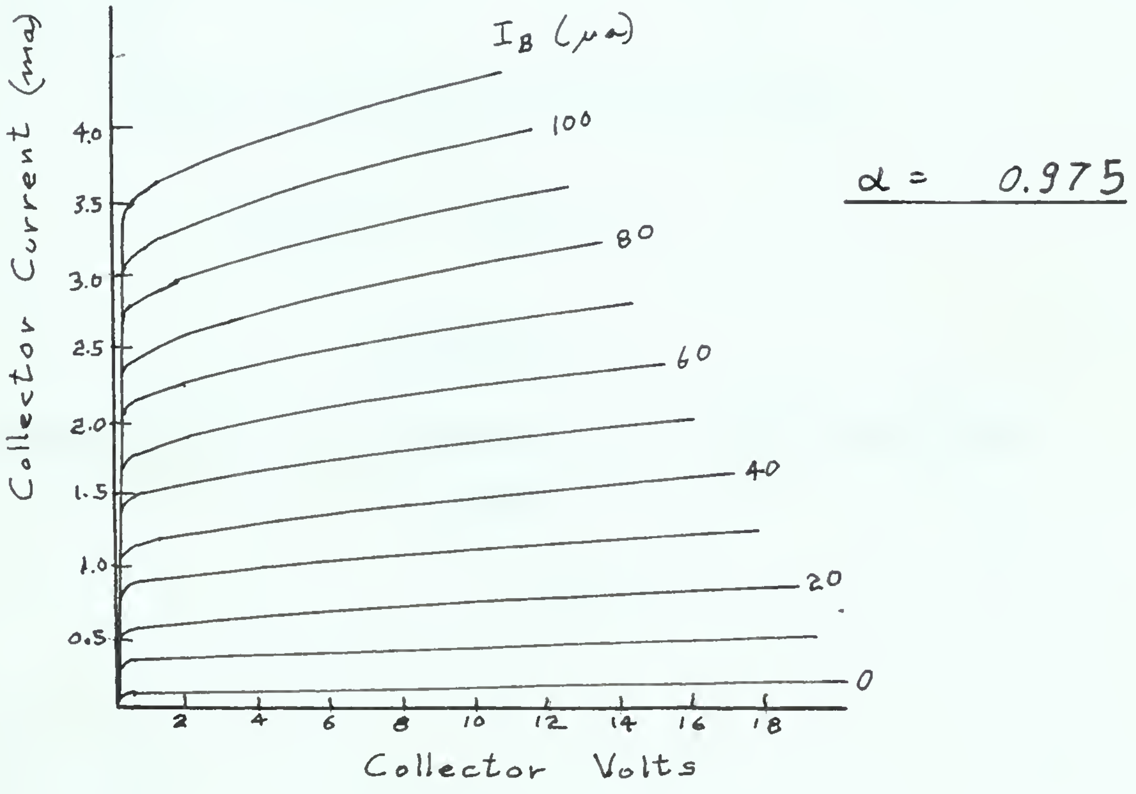

FIGURE B.1

COMMON-EMITTER CHARACTERISTICS OF TRANSISTOR

2N43A

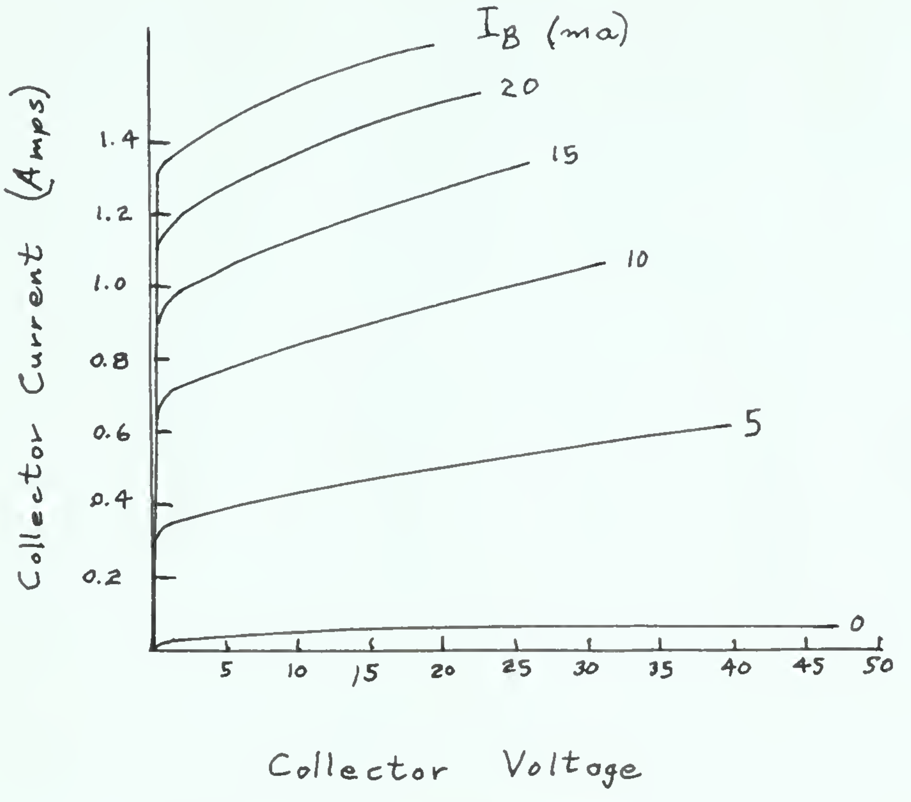

FIGURE B.2

COMMON-EMITTER CHARACTERISTICS OF TRANSISTOR

2N540A

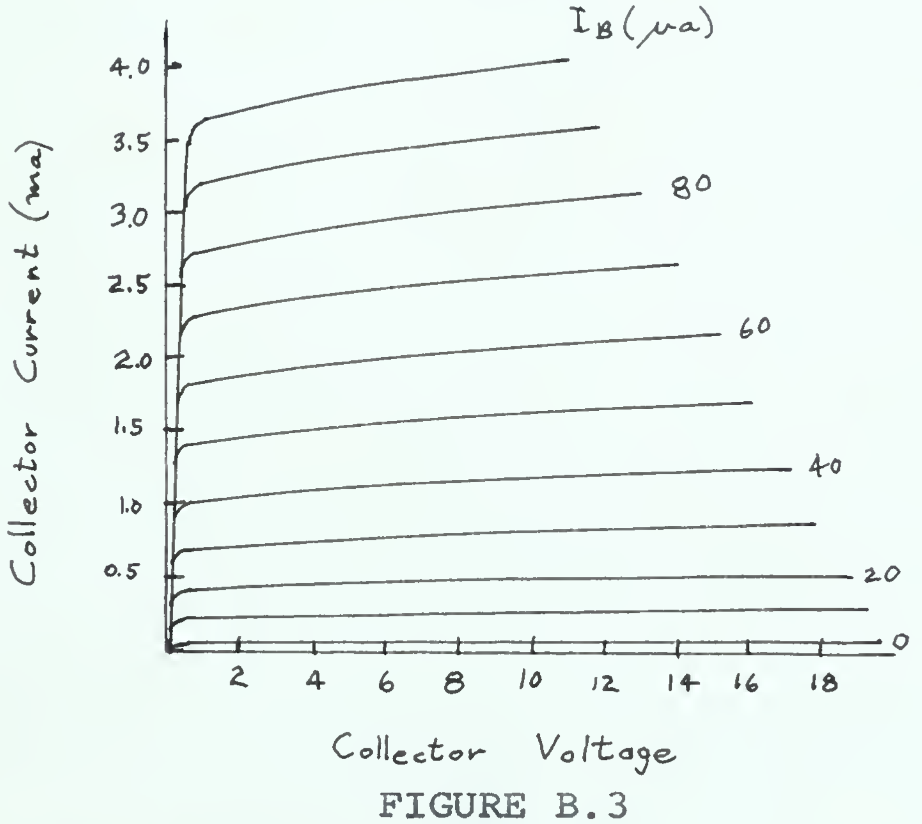

FIGURE B.3

COMMON-EMITTER CHARACTERISTICS OF TRANSISTOR

2N118